MARCH • 1955

Proceedings OF THE IRE

55 SESSIONS
AND SYMPOSIA

704 ENGINEERING EXHIBITS

ANNUAL BANQUET

COCKTAIL PARTY

ANNUAL MEETING

WOMEN'S PROGRAM

1955 IRE

NATIONAL CONVENTION AND ENGINEERING SHOW

MARCH 21–24 • NEW YORK
WALDORF ASTORIA HOTEL
AND KINGSBRIDGE ARMORY

CONVENTION PROGRAM, PAGE 347; WHAT TO SEE AT THE SHOW, PAGE 1A; TABLE OF CONTENTS FOLLOWS PAGE 128A.

The Institute of Radio Engineers

An Important Announcement to Industry

SILICON *POWER* RECTIFIERS

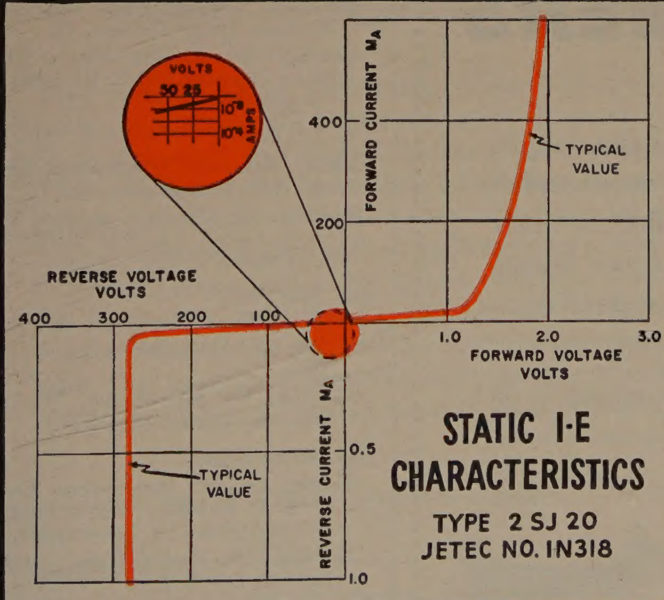
AVAILABLE FOR THE FIRST TIME IN PRODUCTION QUANTITIES

PRECISION ELECTRICAL EQUIPMENT
Bogue
SINCE 1892

★ These units are ideally suited for aircraft and guided missile requirements. Other typical applications that can benefit from their superior characteristics are power rectifiers in commercial equipment, magnetic amplifiers, clipping, meter protection and counter circuits. Anxiety over temperatures is completely eliminated when they are used in digital computers. Automation and control engineering suggest additional fields.

Performance:

1. Rectification Efficiency Over 99%
2. Forward Voltage Drops Averaging 1.5 Volts at 200 MA
3. Peak Inverse Voltages to 1,000 Volts
4. Operates Continuously up to 200°C.
5. Leakage Current as Low as 10-10 amperes
6. Rectification Ratios as High as 10°
7. Practically Flat Zener Characteristics



STATIC I-E CHARACTERISTICS
TYPE 2 SJ 20
JETEC NO. IN318

Characteristics:

1. HIGHEST EFFICIENCY
2. HIGH CURRENT
3. HIGH VOLTAGE
4. HIGH AMBIENT OPERATION
5. HERMETICALLY SEALED
6. SMALL IN SIZE
7. LIGHT IN WEIGHT
8. RUGGED—ALL WELDED
9. LOW FORWARD DROP
10. LOW LEAKAGE

Write for fully illustrated and informative Bulletin SR-18-3

Jetec No.	TYPE	Forward Drop @ 200 MA	Forward Current Continuous	Power Current Peak	Peak Inverse
IN 316	2SJ5	2V Max	200 MA	2A	50V
IN 317	2SJ10	2V Max	200 MA	2A	100V
IN 318	2SJ20	2V Max	200 MA	2A	200V
IN 319	2SJ30	2V Max	200 MA	2A	350V
IN 320	2SJ50	2V Max	200 MA	2A	500V

Units with peak inverse rating of 850 volts available in sample quantities.

BOGUE

BOGUE ELECTRIC MANUFACTURING COMPANY
PATERSON 3, NEW JERSEY

Meetings with Exhibits

As a service both to Members and the industry, we will endeavor to record in this column each month those meetings of IRE, its sections and professional groups *which include exhibits.*

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March 21-24, 1955
Radio Engineering Show and I.R.E. National Convention, Kingsbridge Armory and Kingsbridge Palace, N.Y.C.
Exhibits: Mr. William C. Copp, Institute of Radio Engineers, 1475 Broadway, New York 36, N.Y.

April 15-16, 1955
Ninth Annual Spring Technical Conference, Cincinnati Section, IRE, Engineering Society of Cincinnati Bldg., Cincinnati, Ohio
Exhibits: Mr. Clyde G. Haehnle, Crosley Broadcasting Corp., 140 West Ninth St., Cincinnati 2, Ohio

April 27-29, 1955
Seventh Regional Technical Conference & Trade Show, Hotel Westward Ho, Phoenix, Ariz.
Exhibits: Mr. George McClarathan, 509 East San Juan Cove, Phoenix, Ariz.

May 9-11, 1955
National Conference on Aeronautical Electronics, Biltmore Hotel, Dayton, Ohio.
Exhibits: Mr. William Klein, 1472 Earlham Drive, Dayton, Ohio

May 18-20, 1955
National Telemetering Conference, Morrison Hotel, Chicago, Ill.
Exhibits: Mr. Kipling Adams, General Radio Company, 920 S. Michigan Ave., Chicago, Ill.

June 2-3, 1955
I.R.E. Materials Symposium, Convention Hall, Philadelphia, Pa.
Exhibits: Mr. Merritt A. Rudner, United States Gasket Co., 611 North Tenth St., Camden 1, N.J.

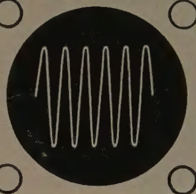
Aug. 24-26, 1955
Western Electronic Show & Convention, Civic Auditorium, San Francisco, Calif.
Exhibits: Mr. Mal Mobley, 344 N. La-Brea, Los Angeles 36, Calif.

Sept. 12-16, 1955
Tenth Annual Instrument Conference & Exhibit, Shrine Exposition Hall & Auditorium, Los Angeles, Calif.
Exhibits: Mr. Fred J. Tabery, 3443 So. Hill St., Los Angeles 7, Calif.

Note on Professional Group Meetings: Some of the Professional Groups conduct meetings at which there are exhibits. Working committeemen on these groups are asked to send advance data to this column for publicity information. You may address these notices to the Advertising Department, and of course listings are free to IRE Professional Groups.

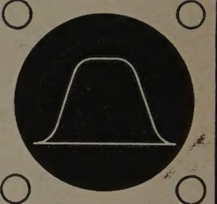
New! MULTI-PURPOSE
Sweep Signal Generator

4.5 to 120 mc.



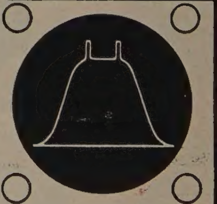
CW Output
4.5 to 120 MC
0.1 to 300,000 μv

AM Modulated CW
30% at 1000 cps.

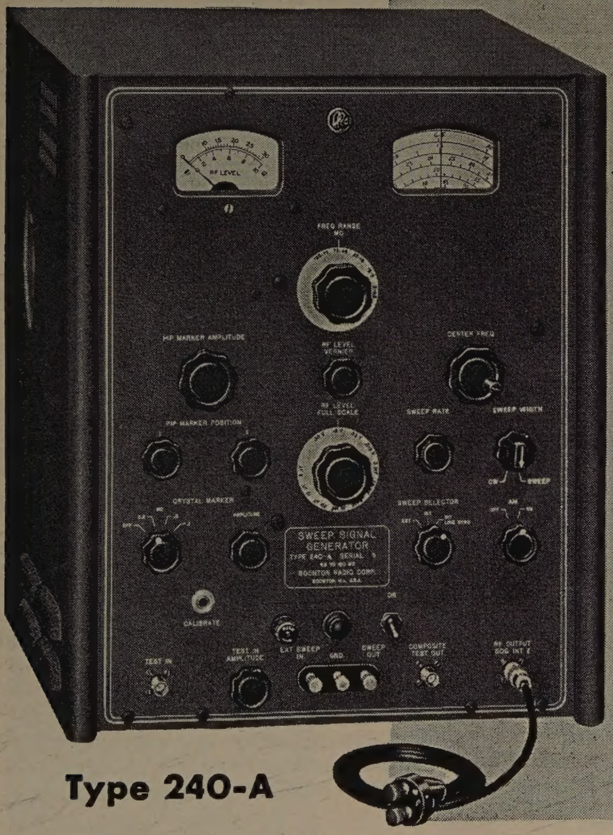


Broad Band Sweep
±1% to ±30% of
Center Frequency

Crystal Controlled
Identification of
Frequency—0.1, 0.5,
2.5 MC Spacing



Interpolation Pips
Mark Frequencies
From Crystal
Identification



Type 240-A

- A wide range of continuously variable linear-frequency sweep widths.
- Crystal controlled frequency identification.
- Adjustable frequency-interpolation pip marks.
- Internal mixer for forming composite display signal.
- Accurate continuously-tuned CW with choice of internal AM.
- Internal crystal calibrator for CW.
- Wide range of calibrated output voltage.

The Sweep Signal Generator Type 240-A is a continuously-tuned, accurate CW Signal Generator with internal AM. The output voltage is continuously monitored, indicated and calibrated over a wide range. The CW Signal can be calibrated against an internal crystal. Electronic sweep circuits are included which produce an output signal having an AGC-controlled, constant-amplitude and a wide range of continuously-variable linear-frequency sweep widths. Two systems are included for frequency identification while sweeping. One of these is crystal controlled and the other is an interpolation system. An internal mixer adds the frequency identification information to the test receiver output signal prior to its connection to the display oscilloscope.

SPECIFICATIONS:

RF FREQUENCY RANGE: 4.5 to 120 MC continuously variable in five ranges.

RF FREQUENCY ACCURACY: ±1%.

RF OUTPUT VOLTAGE: 1 to 300,000 microvolts. 0.1 to 30,000 microvolts with external attenuator.

AMPLITUDE MODULATION: Factory adjusted to 30% from internal 1000 cps oscillator.

RANGE OF SWEEP WIDTHS: Continuously variable from ±1% of center frequency to ±15 MC or ±30% of center frequency whichever is smaller.

LINEARITY OF SWEPT RF FREQUENCY: Within 10% over middle 80% of sweep excursion, within 20% over remainder.

FLATNESS OF SWEPT RF OUTPUT: Within 7% under all conditions.

FREQUENCY IDENTIFICATION MARKS: Crystal frequency identification spaced 0.1, 0.5, 2.5 MC. Tuning dial identifies center mark. Two adjustable-position interpolation pip marks.

PRICE: $1375.00 FOB BOONTON, N. J.

BOONTON RADIO *Corporation*
BOONTON·N·J·U·S·A·

Booths 225-227 Instruments Ave., Radio Engineering Show

News-New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 42A)

Graphic Recorder

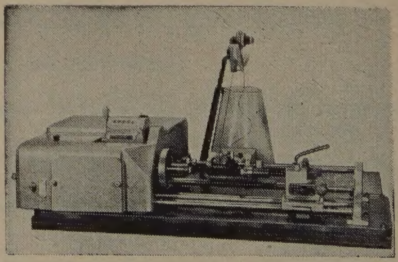
Century Electronics Co., P.O. Box 648, Cocoa, Fla., has developed a direct writing recorder which was especially designed for the recording of low frequency telemetering data; however, applications have also been found in monitoring guided missile firing sequences, computers, etc. The standard telemetering recorder has four chart speeds compatible with data from commutated and low frequency continuous channels. Usable response can be obtained up to 100 cps with a Century compensated amplifier. Any number of chart speeds up to a maximum of eight may be furnished on special units.

The standard recorder includes three center-tapped pen motors and one dual polarity timing and event marker. When voltage sensitive paper is used, static data pens may be employed to record on-off functions. For some applications, a combination of dynamic and static marking pens are provided.

The recorder has adjustable static marking pens, and dual polarity timing-event marker. It is ($8\frac{3}{4}\times8\frac{1}{2}\times16$ inches in size). Models are available for voltage sensitive or heat sensitive chart paper. Information is plotted in rectilinear coordinates. These recorders are designed for multi-unit installation in standard racks, and a 30-channel recording system may be installed in approximately 48 inches of a standard rack.

Bobbin Winder

Geo. Stevens Mfg. Co., Inc., Pulaski Rd., at Peterson, Chicago 30, Ill., is now offering a screw feed bobbin and resistor winder featuring a new finger-tip resettable wire guide. Model 212-AM winds all types of random wound bobbin coils, solenoids, repeater coils, relay coils as well as space wound coils, resistors and distributed constant delay lines from $\frac{1}{8}$ inch to $6\frac{1}{2}$ inches long and up to 4 inches outside diameter. Coils up to $12\frac{1}{2}$ inches long may be wound by specifying Model 212-AML.



Upon completion of the winding cycle, the wire guide carriage can be instantly returned to the identical starting point by finger-tip pressure on the clutch release button. This action completely eliminates split nuts. The wire guide carriage is adaptable to multiple winding.

(Continued on page 48A)

News-New Products

(Continued from page 46A)

Conace and Kellogg Appointed by Sperry

Appointment of several men to key executive positions in the **Sperry Gyroscope Company's** new Aeronautical Equipment Division was announced today by H. C. Bostwick, division manager.

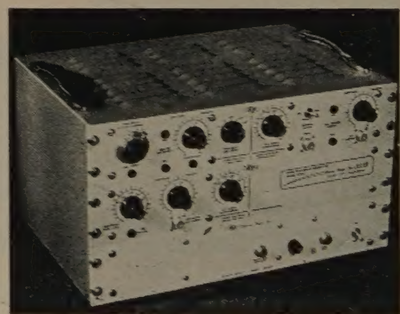
Conace

Kellogg

Sales manager of the new division is Frank Conace, former director of field service engineering. Spencer Kellogg, formerly department head of flight instrument engineering, assumes the responsibilities of engineering manager.

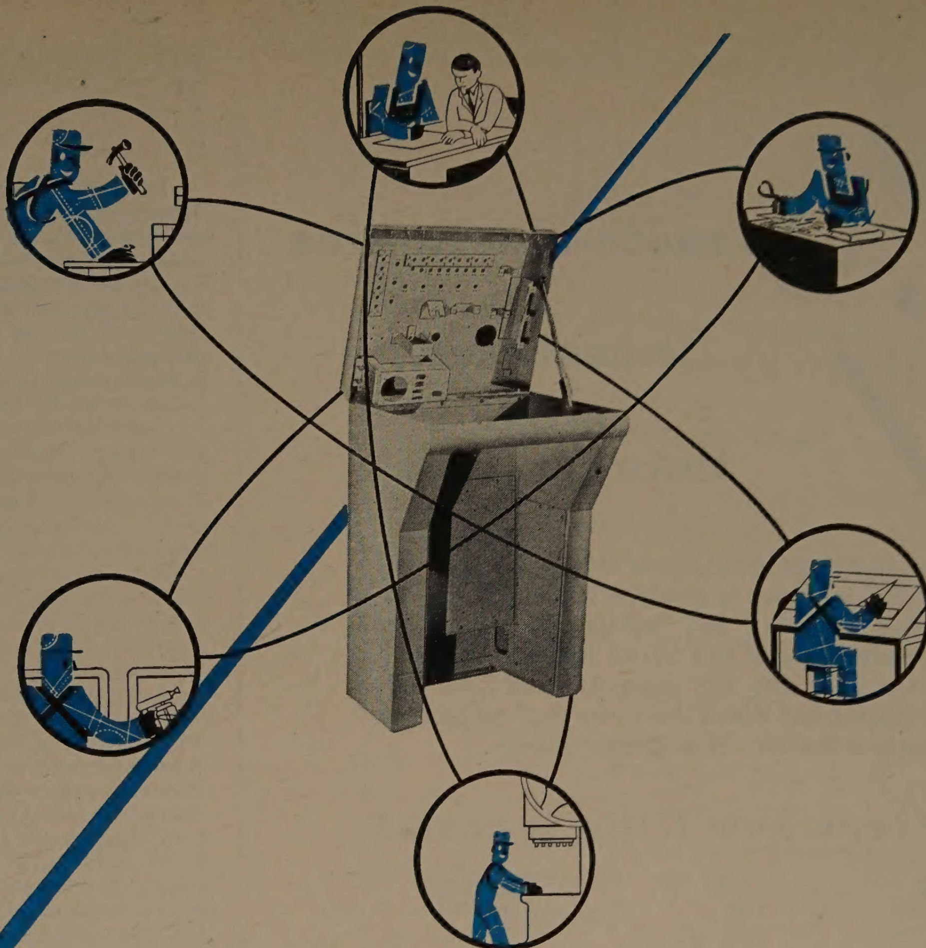
Time Delay Generator

The Model 1310A long time delay generator developed by **Electro-Pulse, Inc.**, 11811 Major St., Culver City, Calif., provides controlled and continuously variable time delays over a range (20 μs to 10 seconds delay).



The instrument is designed for wide use in pulse width and spacing measurements, control of gating circuits, geophysical and biological studies, and other timing applications.

Delay is covered in 5 decade ranges and controlled by a 10 turn potentiometer. Long term accuracy is 1 per cent of full scale for each range (0.5 per cent with standard modification available), and jitter is 0.01 per cent. Blocking oscillator pulses at the reference and delay termination are 0.5 microsecond wide, with an amplitude of at least 45 volts. Positive and negative variable width pulses (10 to 100,000 μs in 4 decade ranges) are also available at the reference and delay points, variable in amplitude to 25 volts maximum. In addition a positive or negative pulse, variable in width from 20 μs to 10 seconds is available.



Industrial Engineering Notes

FCC Actions*

To accommodate the ever-increasing demand for vehicular radio facilities in the domestic public, public safety, industrial and land transportation radio services, the Federal Communications Commission has invited comments by March 28 to its proposed rule making which would reduce channel separations for stations operating in these services in the 25–50 mc and 152–162 mc frequency bands. In addition, the FCC set forth new narrow band frequency stability and emission standards and amortization schedules for existing equipment. The proposal is to amend parts 2, 6, 10, 11 and 16 of the rules so as to list the frequencies in the 25–50 mc band with 20 kc separation (in lieu of the present 40 kc spacing) and also list land mobile service frequencies in the 152–162 mc band with 15 kc separation (reduced from the present 60 kc spacing). "Split channel" assignments in the latter band normally would be made only on an alternate 15 kc basis so as to provide 30 kc separation in the same local service area. However, the commission also invites comments on a 20 kc assignment plan for the 152–162 mc band, with resultant 40 kc separation of local assignments. If information received indicates that the 15 kc plan would not result in the most effective frequency utilization for that band, the 20 kc spacing will be adopted.

Industry Statistics

Fiscal year 1954 marked the twentieth anniversary of the Federal Communications Commission, and with the end of that year over 1.2 million broadcast authorizations in the radio field alone were on the commission's books. These authorizations cover the use of over 700,000 transmitters, it was revealed in the annual report for the fiscal year ending June 30. In the non-broadcast services, the commission reported the following outstanding authorizations at the end of the fiscal year: 46,000 marine stations with 44,000 transmitters, 40,000 aviation stations with 42,000 transmitters, 21,000 industrial stations with 146,000 transmitters, 15,000 public safety stations with 165,000 transmitters, 14,000 land transportation stations with 139,000 transmitters, 123,000 amateur stations with 116,000 transmitters; also over 1,600 common carrier radio stations, and nearly 600 experimental radio stations. In the broadcast field, authorizations neared the 6,000 mark in the following classifications: 2,697 commercial AM stations, 573 commercial TV stations, 30 educational TV stations, 569 commercial FM stations, 123 educational FM stations, 1,728 pickup, studio-transmitter links, and other auxiliary stations. FCC also reported that nearly 850,000 commercial radio operators and more than 120,000 amateur radio operators held permits on June 30.

* The data on which these Notes are based were selected by permission from *Industry Reports*, issues of January 13 and 17, published by the Radio-Electronics-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.

(Continued on page 57A)

Industrial Engineering Notes

(Continued from page 52A)

RETMA ACTIVITIES

A proposal that the Federal Communications Commission set up study groups on various types of spurious radiation with each group comprising FCC and industry engineers and a member of the commission is contained in the comments filed by Dr. W. R. G. Baker, Director of the Engineering Department, on behalf of RETMA on the FCC's proposed spurious radiation order (Docket 9288). Accompanying the general comments filed with the FCC were copies of the reports of seven task forces which have been working on particular aspects of the FCC proposed regulations. Dr. Baker commented as follows on these reports: "A review of these reports leads to the suggestion that any action taken at this time by the commission on the overall problem of spurious radiation might appropriately proceed to go as far as practical at this time and then put in place a definite continuing program to deal promptly and effectively with those areas as to which specific limits cannot now be established." Dr. Baker added that "there are areas in which task forces have found it necessary to conclude that specific recommendations are not possible at this time." The RETMA comments revealed that the task forces had encountered a relatively new phase of the spurious radiation problem involving the radiation characteristics of foreign broadcast receivers imported into this country. "The preliminary indication," Dr. Baker said, "is that several types of these receivers are of such a design as to create an unusually severe problem with respect to meeting any of the limits which have been under consideration in this docket and the attached proposals. Any further work done in this field should be organized so as to deal with this specific phase of the problem."

TECHNICAL

A flat, transparent television picture tube has been demonstrated by the Navy at the West Coast Electronics Laboratory of Willys Motors, Inc. The tube reportedly was developed under a Navy contract in connection with a long-range program for simplifying aircraft instruments. According to the Navy announcement, the new TV picture tube, invented by William Ross Aiken, Director of Research at the laboratory, is approximately the size and shape of a metropolitan city telephone book, is three inches in thickness, and consists of a phosphor screen sandwiched between glass plates. The tube functions by electronically exciting selected areas or spots on the phosphor screen. This is accomplished by the following means: an electronic beam is injected along a horizontal edge of the tube; this beam flows in a field-free region along the edge of the phosphor screen and adjacent to a row of transverse deflection plates. Through control of the voltages of these deflection plates, the beam is bent vertically at any desired place along the edge of the tube.

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Industrial Engineering Notes

(Continued from page 57A)

The beam then flows vertically in a second field-free region between a series of transparent deflection plates and the electrically charged phosphor screen. The tube, compact and portable, employs only electrostatic principles, which eliminate the necessity of using conventional magnetic components, which are heavy, costly and require greater electrical power, the announcement stated. Demonstrated to representatives of the American airframe industry, the new TV picture tube is a part of the program to develop an instrument panel consisting of only two basic instruments—both of which would be television picture tubes—mounted atop one another in front of the pilot. One picture would give the data on altitude, speed and attitude of the aircraft and the other would give broad physical features of the earth below and be depicted by analogy. The Navy expects that the first experimental aircraft using this instrument and control system will be flown about 1958. . . . **In order to provide a direct, continuing link between the National Bureau of Standards and the organized science and technology of the nation, 12 technical area advisory committees, including several related to electronics, have been established in the bureau during the past year.** The committees represent the society from which they are drawn rather than committees of the bureau in order to encourage objectivity in their functions. The committees, which will supplement the Bureau's Statutory Visiting Committee, have been set up as a result of recommendations made by the *ad hoc* Evaluation Committee appointed by the Secretary of Commerce in April 1953 to evaluate the bureau's program in relation to national needs. Under the chairmanship of Dr. Mervin J. Kelly, President of Bell Telephone Laboratories, the Evaluation Committee conducted a comprehensive survey and reported in October 1953 that the bureau's statutory functions were well conceived and its operations generally sound. At the same time, it was pointed out, the committee recognized the desirability of some means whereby the needs of the nation's scientific and engineering societies could be expressed and transmitted to the bureau for implementation in its program. The American Institute of Electrical Engineers has appointed an advisory committee, under the chairmanship of Dr. Ralph Bown of Bell Telephone Laboratories, to consider programs in electricity and electronics. Work in electricity is primarily concerned with development and improvement of standards and methods of measurement for all electrical quantities. Research in electronics at NBS is directed toward obtaining knowledge of basic electronic phenomena and the properties of materials of potential significance to electronics. Also included on this committee

(Continued on page 61A)

Industrial Engineering Notes

(Continued from page 59A)

are: Dr. C. G. Suits, General Electric Co.; Dean F. E. Terman, Stanford University; Dr. E. W. Engstrom, Radio Corp. of America; Robert C. Sprague, Sprague Electric Co., and Dr. J. A. Hutcheson, Westinghouse Electric Corp. Also, an advisory committee appointed by the IRE has as its principal interest the NBS Central Radio Propagation Laboratory in Boulder, Colo. Chairman of this group is Dr. A. W. Straiton, Electrical Engineering Research Laboratory, University of Texas. Also serving on the committee are: Dean F. E. Terman, Stanford University; Prof. Henry G. Booker, Cornell University; Harold O. Peterson, Radio Corp. of America; Stuart L. Bailey, Jansky & Bailey, and Dean William L. Everitt, University of Illinois. . . . **New efficiency in the control of servomechanisms is claimed for a new Navy-designed "half-cycle response time" magnetic amplifier which is detailed in a research report recently made available to industry through the Commerce Department's Office of Technical Services.** At the same time, the OTS announced another Navy-developed unit which accurately and continuously checks the degree of impedance match between an antenna and its transmission line. The report can be ordered by number PB 111442, "A New Magnetic Servo Amplifier," from the Office of Technical Services, Commerce Department, Washington 25, D. C., for $1 each. The second report can be ordered by number PB 111439, "Impedance-Match Indicating Devices," from the Office of Technical Services, Commerce Department, Washington 25, D. C., for $1 each.

Television

The Federal Communications Commission has authorized a new experimental TV station to test a low-cost operation, on a non-profit basis, at Manson, Wash. The Manson Community Television Co., formed for that purpose, was authorized to experiment in converting the signals of distant VHF stations, amplifying and retransmitting them to that area on UHF Channel 16 with effective radiated power of 200 watts from various experimental types of antennas not to exceed 30 feet above the ground. The experiment, according to the FCC announcement, will be on a non-profit basis and subject to certain engineering conditions. Rules have been waived insofar as they require an operator to be on duty at the transmitter, but an operator is required for observations and adjustments. The call letters will be transmitted in International Morse Code automatically at the beginning and end of each period of operation and at least once during each operating hour. Because of its purpose, the hours of operation are not limited.

Professional Group Meetings

AERONAUTICAL AND NAVIGATIONAL ELECTRONICS

The New York Chapter of the Professional Group on Aeronautical and Navigational Electronics met on December 16 with R. J. Bibbero presiding. David S. Little spoke on the electronic facilities of American Airlines at LaGuardia Field in New York. Afterwards, there was a tour which included the reservation input stations to Reservisor computer, the magnetic drum reservisor, automatic teletype, tower GCA, and airborne communications.

ANTENNAS AND PROPAGATION

On October 22 at the Western Society of Engineers Auditorium, the Chicago Chapter of the Professional Group on Antennas and Propagation met. James Eakins, a development engineer at Motorola, presented a paper entitled "Propagation Loss Measurements at 450 mc Applied to Mobile Systems Engineering."

The Philadelphia Chapter met on November 17 with R. P. Schwartz presiding. E. I. Hawthorne, University of Pennsylvania, was the speaker. In his paper, "Shielding with Coated Glass on Thin Conductive Surfaces," he presented a mathematical analysis of the transmission of electromagnetic fields through thin conducting planes, infinite in extent. In particular, he considered two sources: a magnetic and an electric current element assumed normal to the sheet. The field along the axis of the source on the opposite side of the sheet is determined and compared with the field which would exist if the sheet were not present.

AUDIO

The Philadelphia Chapter of the Professional Group on Audio met on October 28 at the WCAU Studio. H. E. Roys was the presiding officer. Two papers were presented. J. E. Volkmann, of RCA, spoke on "Stero-Acoustic Principles." He discussed the basic acoustic principles involved in the recording and reproduction of stereophonic sound. Other types of multi-channel sound systems and loudspeaker placement for spatial sound effects were also described. R. C. Moyer, also of RCA in Camden, delivered a paper entitled "High-Fidelity Pre-Recorded Tapes." He reviewed the equipment and techniques which have been developed for the production of high-fidelity pre-recorded tapes. Demonstrating both types, Mr. Moyer described monaural half-trach tapes and dual-trach stereophonic recordings.

At the Lovelace Clinic on November 23 the Albuquerque-Los Alamos Chapter met. Frank McIntosh, of McIntosh Corporation, was the speaker. Mr. McIntosh described and demonstrated the new McIntosh 200 watt audio amplifier.

Harold Mull presided at the meeting of the Cleveland Chapter on November 29.

(Continued on page 68A)

From Original Engineering...

From Original Engineering...and from its half-century of specialized skills and experience, Machlett once again creates new riteria of electron tube performance for high power electronics.

Iachlett is first again to offer significant, original esign for high power triodes. It was six years go that Machlett perfected a heavy duty power ıbe series—the first from any manufacturer. ccepted at once by equipment manufacturers, ow universally used, these tubes were also *aid the compliment of imitation.*

Vith the New Machlett Triodes
COSTS GO DOWN

ecause filament power is reduced—
p to 60% (enough, in many cases, to pay for the tube).

ecause tube life is longer—
ıcreases of over 100%, compared to conventional types.

ecause maintenance is cut—
eplacement minimized; cleaning simplified.

ecause handling costs are low—
eight down 60 to 70% for air cooled types.

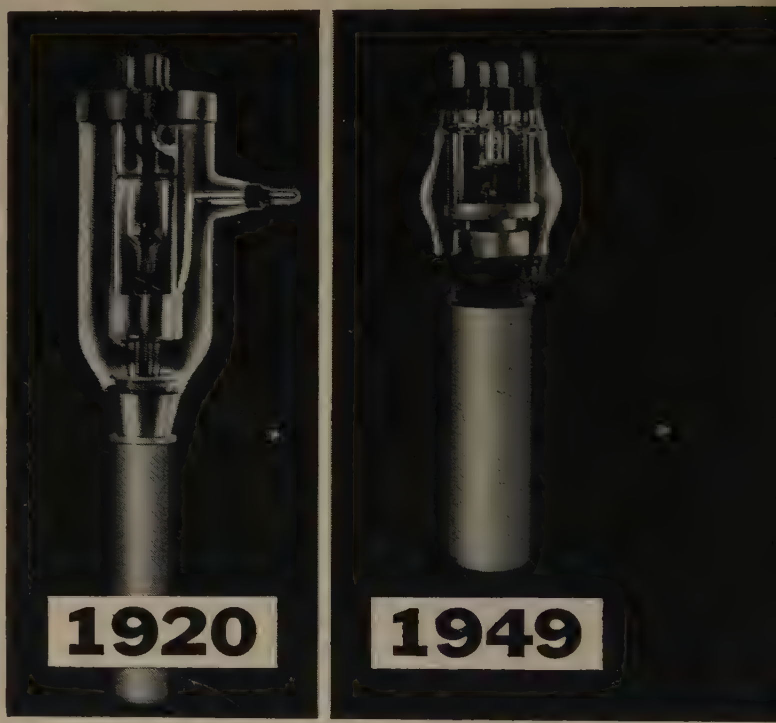
Vith the New Machlett Triodes
PERFORMANCE GOES UP

ecause plate and grid ratings are higher—
roader range of operation is possible.

ecause safety margins are usefully increased—
or thermal, mechanical and electrical ratings.

ecause lead inductances are very low—
rcuit parasitics are reduced by as much as 10-to-1.

ecause transconductance is high—
late efficiency is increased, grid drive reduced, able performance assured over broadest loading range.



The above tubes portray progressive evolution in electron tube design. Left, Type 892, uses long, high inductance electrode leads and large glass envelope—a design now over 35 years old. Center, Machlett developed, electrical equivalent, industrial ML-5668, has stronger, less inductive internal structures, short glass envelope and sturdier seals, thicker anode with double heat dissipation capacity. Right, most modern tube, Machlett's new ML-6422, uses cylindrical electrode supports for lowest inductance, great stability; large contact area terminals for great seal strength; close-spaced, precisely-aligned electrodes for low drive and high efficiency; thick-wall anode for cool tube operation, high overload capacity; stress-free thoriated-tungsten filament for high load current, low heating power, and longer life.

MACHLETT

And Your First Cost is Only One Postcard—that's all it takes to write Machlett Laboratories for the full story of these premium design, rugged, coaxial triodes. Learn now what to expect of a modern power triode.

MACHLETT LABORATORIES, INC.
1063 Hope Street, Springdale, Connecticut

Professional Group Meetings

(*Continued from page 63A*)

"An Experiment in Stereophonics" was the title of the paper delivered by S. J. Begun of the Clevite-Brush Development Company.

On November 19 at the Western Society of Engineering Building the Chicago Chapter met. D. E. Weigand, Senior Physicist at Armour Research Foundation, spoke on "A Flux Sensitive Head for Recording and Playback."

Broadcast and TV Receivers

With High Christian presiding, the Chicago Chapter of the Professional Group on Broadcast and TV Receivers met on November 19. Samuel Tarantur of Admiral Corporation spoke on "An Encoder for the NTSC Color Signal."

On October 1 the Chicago Chapter met again at the Western Society of Engineers Building. Jack Bridges, of Zenith Radio Corporation, delivered a paper called "The Second Detector—A Determinant of Fringe Area Performance."

Broadcast Transmission Systems

On October 1 at WBKB Studios, the Chicago Chapter of the Professional Group on Broadcast Transmission Systems met. G. M. Ives presided. Forming twelve groups, the chapter made a tour of facilities at WBKB Studios.

Circuit Theory

The Los Angeles Chapter of the Professional Group on Circuit Theory met on January 13. J. L. Barnes, Ramo-Wooldridge Corporation, spoke on "Three Dimensional Network Duality," and L. Weinberg, of Hughes Aircraft Company, discussed "Methods of Network Synthesis."

With E. Mittman presiding, the Chicago Chapter met on October 1 at the Western Society of Engineers Building. B. S. Parmet, of Motorola, spoke on "Combined Plate and Cathode Video Compensation Using Laplace Transforms."

On October 22 the Chicago Chapter again met. Bernard S. Parmet presided. "Noise Sources in High Frequency Tubes," was the paper delivered by Robert Adler, director of Research at Zenith Radio Corporation.

The Philadelphia Chapter met at Towne School of the University of Pennsylvania on November 9. Sidney Darlington of Bell Telephone Laboratories spoke on "Network Synthesis Using Tchebycheff Polynomials." Dr. Darling reviewed existing methods for the solution of the interpolation problem of network synthesis, assuming four-pole network realization. After discussing the relative merits of the procedures, he gave a detailed review of his work on network synthesis using Tchebycheff polynomials.

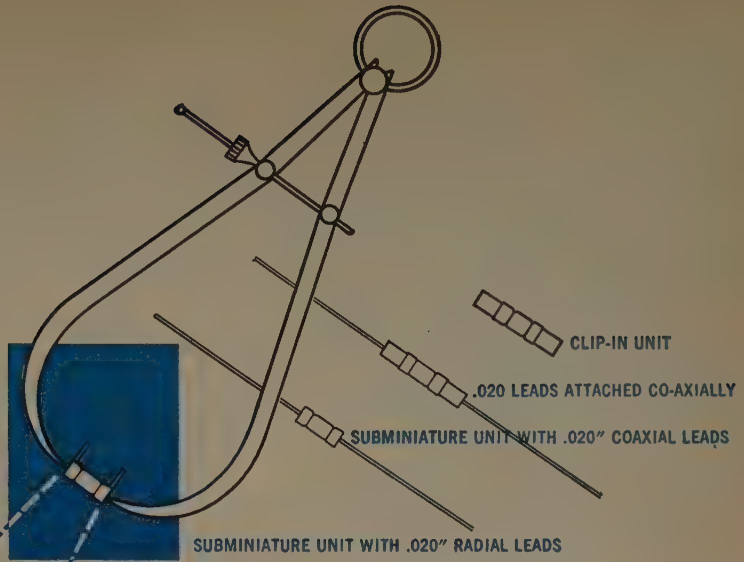
On October 12 three papers were deliv-

(*Continued on page 72A*)

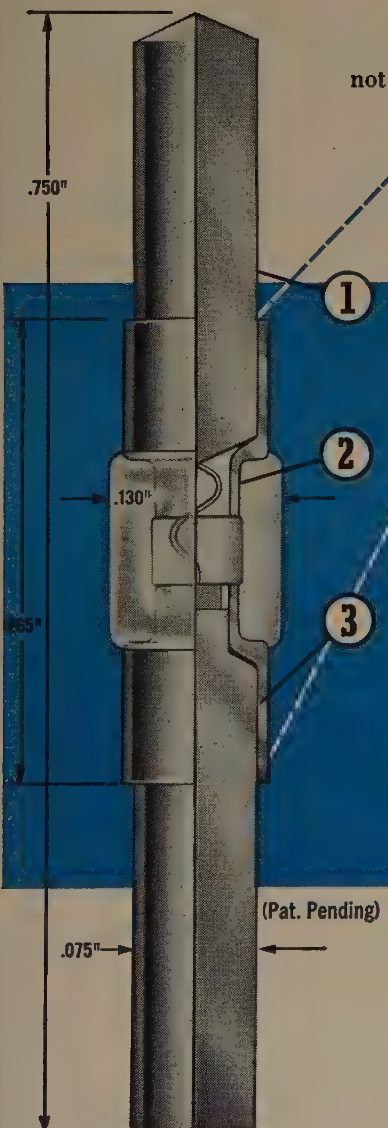
NOW AVAILABLE

- **GERMANIUM GOLD BONDED DIODES**
- **SILICON JUNCTION DIODES**

in the new PSI diode package . . . designed around user requirements



PSI's revolutionary new package, with advantages not found in any other commercially available diodes, was designed only after an exhaustive survey of user requirements. Space limitations, environmental demands, even assembly procedures became factors in the final design. The result: diodes with demonstrably superior performance, greater versatility, top all-around utility.



(Pat. Pending)

*CHECK THESE BENEFITS . . .

1. VERSATILE LEAD ARRANGEMENT . . . for maximum adaptability, diodes may be obtained in a variety of configurations.

2. GLASS-TO-METAL SEAL . . . for positive moisture resistance, PSI uses a true fusion seal.

3. WELDED CONSTRUCTION . . . for greater strength and freedom from contamination; no low melting point solders are used.

and **your** net benefit from all these features . . .

NEW STANDARDS OF RELIABILITY AND STABILITY

Typical PSI Gold Bonded Diode Characteristics @ 25°C

Forward Current @ 1v (ma)	Inverse Current (μa)	Inverse Working Voltage (volts)
100	100 (−20v)	35
35	10 (−50v)	80
15	25 (−50v) 200 (−200v)	220

Typical PSI Silicon Junction Diode Characteristics

E_s/E_t (volts)	Forward Current @ 1v (ma)	Back Current	
		at 25°C	at 150°C
30/29	80	.01μa (−15v)	5μa (−15v)
55/53	40	.01μa (−30v)	5μa (−30v)
150/145	15	.01μa (−75v)	5μa (−75v)
300/290	5	.01μa (−150v)	5μa (−150v)

a: The saturation voltage (E_s) is measured at 500μa; the transition voltage (E_t) is measured at 20μa.

b: Recovery time: after switching from 5ma forward current to −40v for all these types, back resistance reaches or exceeds 50K in 1μsec.

For complete product specifications, address inquiries to Dept. S-2

PACIFIC SEMICONDUCTORS, INC.

10451 WEST JEFFERSON BOULEVARD

CULVER CITY, CALIFORNIA

Professional Group Meetings

(Continued from page 68A)

ered to the Philadelphia Chapter. A. Papoulis of BPI spoke on "The Displacement of Poles and Zeroes Due to Parasitic Circuit Elements," D. Sunstein of Philco spoke on "A New Self-Oscillating Frequency Converter," and H. J. Woll of RCA discussed "An Analog Multiplier."

Component Parts

The Philadelphia Chapter of the Professional Group on Component Parts met on December 4 at the Engineers' Club. Dr. Leopold Pessel of RCA spoke on "Solderability." He emphasized the chemical and metallurgical factors affecting the ability of metal surface to form effective joints with solders; a test method to evaluate solderability of surfaces on a numerical basis; the outlining of an industry-wide solderability standard, of particular value in improving the reliability of electronic equipment. A question period followed Dr. Pessel's paper.

Electron Devices

At PEPCO Auditorium on December 1 the Washington Chapter of the Professional Group on Electron Devices met. J. Howard Wright presided. A. W. Holt of the U. S. National Bureau of Standards delivered a paper, "The Diode Amplifier." After discussion of the paper and refreshments, there was a short business session at which the following officers for the new chapter were elected: Chairman, J. Howard Wright; Vice-Chairman, H. D. Arnett; Secretary, Harold J. Peake.

The New York Chapter met on November 18 at the General Electric Auditorium. A. K. King was the presiding officer. James M. Early of Bell Telephone Laboratories spoke on "Junction Transistors at Very High Frequency."

The Philadelphia Chapter met, with B. T. Svihel presiding, on November 23. "Dielectric Amplifiers" was the title of the paper delivered by J. R. Horsch of General Electric Company. Dr. Horsch discussed the hysteresis characteristics of dielectric crystals and the principles of amplification by use of dielectrics. He presented the historical background of dielectric amplifiers and showed the performance curves of typical present-day amplifiers. Future development in this field, he said, is dependent largely upon the demand for capacitors with better hysteresis curves.

Electronic Computers

On November 19, the Chicago Chapter of the Professional Group on Electronic Computers met with I. S. Lerner presiding. W. Flurzeig of the Institute of Air Weapons Research spoke on the "Logic of Computers."

The Chicago Chapter also met on October 22. "Representative Problems in Operations Research" was the title of the

(Continued on page 76A)

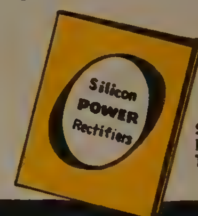
- **Reliability at High Temperatures**
- **High Power Handling Ability**
- **Negligible Leakage Current**
- **No Forward Aging**
- **High Conductance**
- **Miniature Size**
- **Hermetic Sealing**

Transitron's silicon rectifiers meet the long felt need for reliable and efficient power rectification at high temperatures.

Overcoming the basic limitations of selenium, germanium, and the vacuum tube, they provide trouble-free operation over wide ambient temperature ranges.

These rugged rectifiers offer major savings in both size and weight in any high temperature application.

VISIT BOOTH 580 AT THE I.R.E. SHOW, MARCH 21-24



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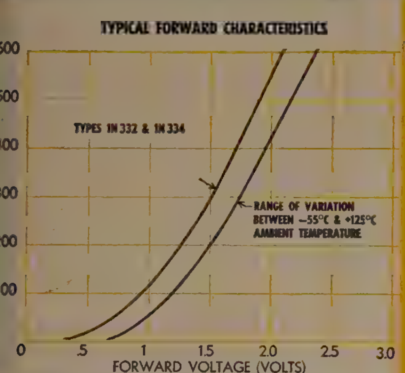
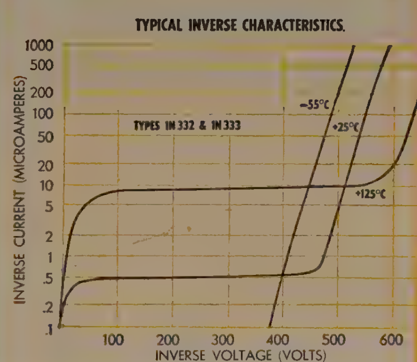
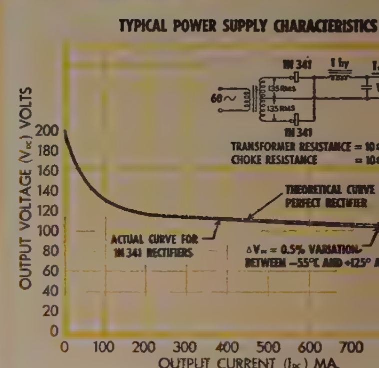






SPECIFICATIONS AND RATINGS AT 125°C

POWER SUPPLY TYPES

TYPE	Peak Recurrent Inverse Voltage (volts)	Maximum Average Forward Current (ma)	Maximum* Average Forward Voltage (volts)	Maximum* Average Inverse Current (ma)
1N341	400	400	3.0	.5
1N342	400	200	3.0	.5
1N343	300	400	3.0	.5
1N344	300	200	3.0	.5
1N345	200	400	2.5	.5
1N346	200	200	2.5	.5
1N347	100	1000	1.0	.5
1N348	100	400	2.5	.5
1N349	100	200	1.0	.5

MAGNETIC AMPLIFIER TYPES

TYPE	Peak Recurrent Inverse Voltage (volts)	Maximum Average Forward Current (ma)	Maximum* Average Forward Voltage (volts)	Maximum* Average Inverse Current (ma)
1N332	400	400	1.5	.10
1N333	400	200	3.0	.10
1N334	300	400	1.5	.10
1N335	300	200	3.0	.10
1N336	200	400	1.3	.05
1N337	200	200	2.5	.05
1N338	100	1000	1.0	.10
1N339	100	400	1.3	.05
1N340	100	200	2.5	.05

Special silicon rectifiers, or combinations, are available to meet specific requirements. Your inquiries are invited.

**Averaged over one cycle for full wave choke input circuit with Rectifier operating at full rated forward current.*

Professional Group Meetings

(Continued from page 72A)

paper delivered by Allan Butterworth, Assistant to the Director of the Institute of Air Weapons Research.

The Los Angeles Chapter met on November 18 and Willis Ware was the presiding officer. Erwin Tomash, Remington Rand West Coast Representative, spoke to the group on "DS-63, A New Business Data Processing Machine." "Patent Law In Engineering" was the name of the paper delivered by Samuel Lindenbert.

On October 27 the Detroit Chapter met at the General Motors Research Center. Roy Hock was the presiding officer. D. E. Hart, General Motors Research, spoke on the computing facilities at the center. He also demonstrated the IBM 701.

At the Franklin Institute on November 16 the Philadelphia Chapter met. Bernard M. Gordon, Vice-President of EPSCO, delivered a paper entitled, "Process Control with Operational Digital Techniques." Mr. Gordon explained how digital-analog simulators combine the advantages of digital precision, analog simplicity and flexibility.

The Philadelphia Chapter also met on October 19; T. H. Bonn was the presiding officer. There were two speakers. A. W. Vance, of the Sarnoff Research Center at RCA, spoke on "Development of a New Large-Scale Three Dimensional Analog Simulator." "Digital Computers for Real Time Simulation" was the subject discussed by Morris Rubinoff of the Moore School of Electrical Engineering.

Engineering Management

The Chicago Chapter of the Professional Group on Engineering Management met on November 19 with Marvin Quatman presiding. "What Keeps an Engineer Out of Management" was the subject discussed by Charles W. Blahna of Motorola.

On December 3 the Philadelphia Chapter met. Irven Travis of Burroughs Corporation spoke to the group on "New Product Planning and the Implementation of New Product Goals as They Affect Engineering Management." Dr. Travis presented the engineering management's problems in new product planning, emphasizing the need to have engineers understand the sales viewpoint and the need to express engineering viewpoints to non-engineers as lucidly as possible.

The Washington Chapter met on November 22 at the National Science Foundation. Karl Keller, Assistant Manager of Westinghouse Electric Corporation, delivered a paper on "Control of Inventory," and Robert Cheek discussed "Decentralized Project Engineering Management."

J. E. Stiles, Vice-Chairman of the PGIE, presided at the meeting of the Chicago Chapter. Joseph Gardberg, of the Teletype Corporation, spoke on "An Electronic Scale Control System."

(Continued on page 77A)

Professional Group Meetings

(Continued from page 76A)

Information Theory

The Albuquerque-Los Alamos Chapter of the Professional Group on Information Theory met on December 8 with John McLay presiding. C. H. Bidwell was the speaker; he delivered a paper called "Signal To Noise Problems in L-3 Coax Development."

Instrumentation

On October 1 the Chicago Chapter met at the Western Society of Engineering Building. F. L. Moseley, President of F. L. Moseley Company, delivered a paper entitled "An X-Y Recorder, Curve Follower, and Data Plotter."

Medical Electronics

The Connecticut Valley Chapter of the Professional Group on Medical Electronics met on October 21. John H. Heller, Director of the New England Institute of Medical Research, spoke to the group on "Medical Electronics."

The San Francisco Chapter met on December 2 with Lee B. Lusted presiding. "Medical Aspects of X-Ray Microscopy" was the name of the paper presented by E. M. Miller, Professor of Radiology at the University of California Medical School. H. H. Pattee and Paul Kirkpatrick, Research Associate and Professor, respectively, in the Department of Physics at Stanford University, jointly presented "Physical and Electronic Techniques in X-Ray Microscopy."

Microwave Theory and Techniques

On November 17 the Philadelphia Chapter of the Professional Group on Microwave Theory and Techniques met at the Towne Building of the University of Pennsylvania. R. P. Schwartz was the presiding officer. D. R. Crosby of RCA spoke on "Theoretical Gain of Microwave Reflectors." He described the physical behavior of reflectors from the field point of view.

Nuclear Science

On October 27 at Mitchell Hall the Albuquerque Chapter of the Professional Group on Nuclear Science met. R. D. Jones delivered a paper to the group entitled "Control Applications of Proton Magnetic Resonance."

The Connecticut Valley Chapter met on November 23 at the Underwater Sound Laboratory. J. E. Basset presided and T. S. Gray was the speaker. He presented a paper called "Electronic Nuclear Instrumentation."

On October 1 the Chicago Chapter met at the Western Society Building. R. C. Goertz, of Argonne National Laboratories, spoke to the group on "Remote Manipu-

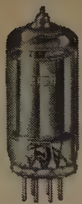
(Continued on page 100A)

FOR HIGH RELIABILITY

THE NEWEST in G.E.'s wide germanium rectifier line is the 1N315. Specifically designed for *high* operating temperatures—up to 85°C.—and for *low* reverse current. (Note: The 1N315A *exceptionally* low-leakage design is also available.) Ideal for use in magnetic amplifiers or other circuits where low-leakage current is important. Write today for details. *General Electric Company, Section X5235, Germanium Products, Electronics Park, Syracuse, N.Y.*

Production quantities are now available!

CHATHAM
HYRATRONS



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Type 2D21W
Similar to 2D21 except ruggedized to conform to MIL-E-1B specifications.

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ce. Wide am-
temperature
nce. Medium
phenolic base
edium metal
Heater 2.5
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Type 323B
Same as 3C23 but has medium shell 5 pin base and small top cap.

pe 2050
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controlled
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mp. Peak in-
anode volt-
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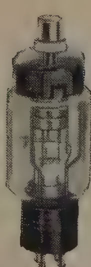
ruggedized to
rm to MIL-E-
pecifications.

Type 394-A
Mercury vapor and Argon filled thyratron for grid controlled rectifier service. Wide ambient temperature tolerance. Heater 2.5 volts, 3.2 amps. Peak inverse anode voltage 1250, aver. anode current 640 ma.



pe 5594
filled thyra-
Operates in
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from −55°
-90°C. Fila-
2.5 volts, 5
Peak inverse
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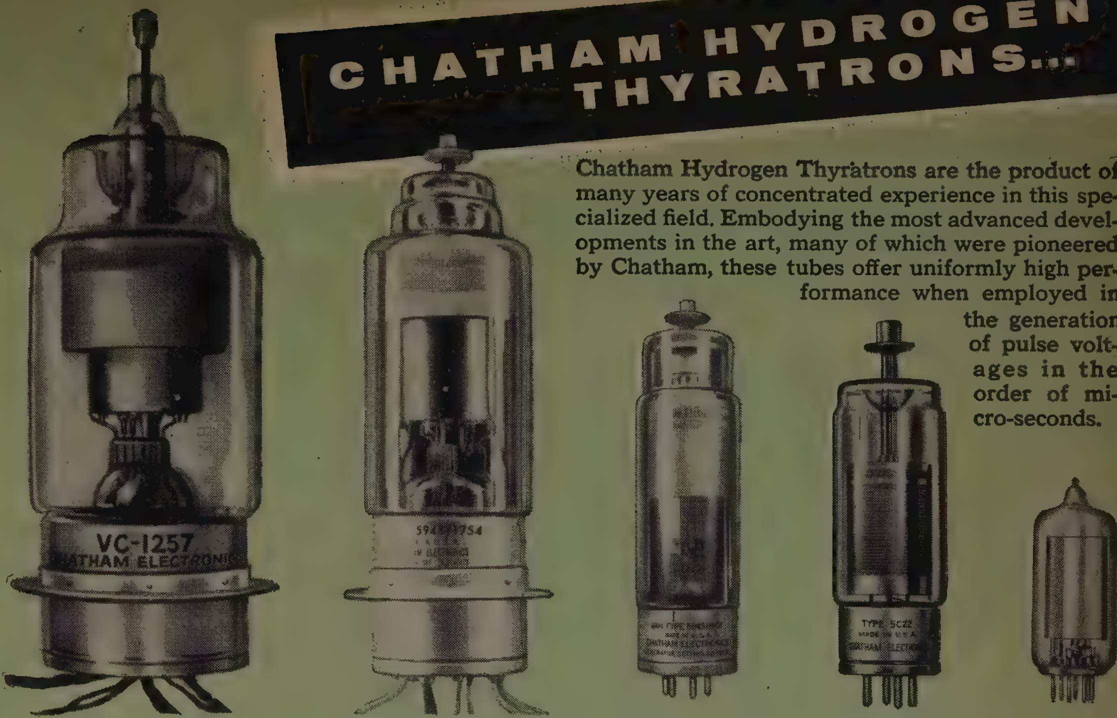
CHATHAM HYDROGEN THYRATRONS...

Chatham Hydrogen Thyratrons are the product of many years of concentrated experience in this specialized field. Embodying the most advanced developments in the art, many of which were pioneered by Chatham, these tubes offer uniformly high performance when employed in the generation of pulse voltages in the order of micro-seconds.



Type VC-1257 Hydrogen filled, zero bias thyratron with hydrogen generator for generation of pulse power up to 33 megawatts.

Type 5948/1754 Hydrogen filled, zero bias thyratron with hydrogen reservoir for generation of peak pulse power up to 12.5 megawatts.

Type 5949/1907 Hydrogen filled, zero bias thyratron with hydrogen reservoir for generation of peak pulse power up to 6.25 megawatts.

Type 5C22 Hydrogen filled, zero bias thyratron with hydrogen reservoir for generation of peak pulse power up to 2.6 megawatts.

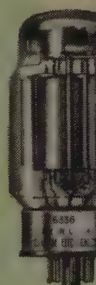
Type VC-1258 Zero bias miniature hydrogen thyratron for the generation of peak pulse power up to 10 KW.

ELECTRICAL DATA*

Type	VC-1258	5C22	5949/1907	5948/1754	VC-1257
Maximum Peak Foward Anode Potential	1000 volts	1600 volts	25000 volts	25000 volts	33000 volts
Maximum Peak Anode Current	20 amps	325 amps	500 amps	1000 amps	2000 amps
Maximum Average Anode Current	0.05 amps	0.200 amps	0.50 amps	1.0 amps	2.0 amps
Maxiumum Heating Factor (epy x prr x ib)	1.0x10^8	3.2x10^9	6.25x10^9	9.0x10^9	20.0x10^9
Nominal Filament Power	12.6 watts	67 watts	95 watts	190 watts	230 watts
Hydrogen Reservoir	No	Yes	Yes	Yes	Yes

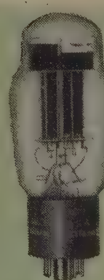
*More detailed information on electrical and mechanical data will be supplied on request.

TWIN POWER TRIODES...

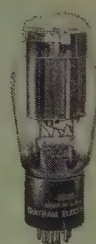


Type 6336
Low mu, high perveance twin power triode for use as a series regulator tube in D.C. power supplies. Plate Dissipation 30 watts per sect.; Amplification Factor 2.7; Plate Resistance 250 ohm per sect.; Plate Current 165-200 ma per sect.; Heater 6.3 volts 4.75 amps total for both sections.

Type 6394
Same as Type 6336 with 26.5 volt 1.25 amp. heater.



Type 6AS7G
Low mu, high perveance twin power triode for use as a regulator tube in D.C. Power Supply Units. Plate dissipation 13 watts per sect.; Amp. Factor 2.0 per sect.; Plate Res. 280 ohms per sect.; Heater 6.3 volts 2.5 amperes total for both sections.



TYPE 6520
Characteristics same as 6AS7G but intended for applications where utmost reliability is required in respect to triode balance, absence of excessive plate current drift and high grid to plate insulation.

Type 884
Argon filled thyratron for use as sweep circuit oscillator in CRT circuits. High stability. Heater 6.3 volts, 0.6 amp. Peak forward anode voltage 300, aver. plate current 75 ma.

Type 885
Similar to Type 884 except heater rating of 2.5 volts, 1.5 amp. Both the 884 and 885 are suitable for grid controlled rectifier applications as well as sweep circuit oscillators.

Type 395-A
A cold cathode thyratron. Requires no filament supply. For grid controlled rectifier and relay applications. Max. D.C. anode voltage 150, Max. anode current 10 ma.

Type 359-A
Cold cathode thyratron. No filament supply needed. Suitable for grid controlled rectifier and relay applications. Max. D.C. anode voltage 180, max. anode current 12 ma.

ELECTRONICS

LIVINGSTO
NEW JERS

Professional Group Meetings

(Continued from page 77A)

lating Equipment for Handling High Level Radioactive Materials." A 16 mm film showing installations and operations was also exhibited.

The Chicago Chapter met again on October 22. Donald H. Loughridge, Dean of the Technological Institute at Northwestern University, presented a paper called "The Present Status of Nuclear Power Development."

"An Up-To-Date Survey of Elementary Particles" was the paper presented by Bernard Hamermesh, Argonne National Laboratory, at the meeting of the Chicago Chapter on November 19. Dr. Hamermesh outlined the known properties of the three additional types of elementary particles discovered in the past seven years.

Quality Control

On October 1 the Chicago Chapter of the Professional Group on Quality Control met. C. Mydill, Director of quality control at Motorola, spoke on supplier-consumer relations. Mr. Mydill outlined the quality control organization at Motorola and its mode of operation.

Vehicular Communications

On December 1 the Detroit Chapter of the Professional Group on Vehicular Communications met at the Engineering Society of Detroit. E. C. Denstaedt, R. Donavon, R. Karber, and J. McFatridge described the Palmer Park OCD Radio Control Center. Added information on the wire line facilities was presented by Sherman Lister of the Michigan Bell Telephone Company. Officers for this year were elected: Chairman, A. B. Buchanan; Vice-Chairman, R. J. DeHaan; Secretary-Treasurer, J. G. Nauman.

Section Meetings

Akron

"Contouring with Electronic Machine-Tool Controls," by N. B. Nichols, Raytheon Manufacturing Company; December 14, 1954.

"The Amplifier as a Link in the Chain of Audio Reproduction," by J. P. Overley, Radio Manufacturing Engineers, Inc.; January 18, 1955.

Albuquerque-Los Alamos

"Application of Radiation Techniques," by Dr. R. S. Wilson, General Electric; December 14, 1954.

Atlanta

Tour of Pan-Electronics Corp., conducted by J. R. Blasingame; December 8, 1954.

Baltimore

"Color Television Broadcasting," by J. L. Hathaway, National Broadcasting Company; January 12, 1955.

(Continued on page 112A)

JK-G3: 10 mc to 150 mc. Miniature size, minimum aging drift, high Q for maximum performance. JK miniaturized "Glasline" crystals meet the growing need for minimum size with maximum stability.

(ACTUAL SIZE)

that unlocks tomorrow's door

CERTIFIED STABILITY

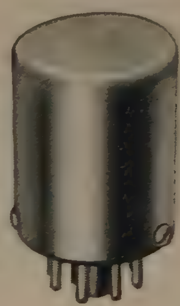
performance is the key to spectrum conservation

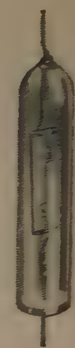
With the radio spectrum already congested, and channels narrowing, frequency control tolerances must narrow too. Equipment that "drifts" or "ages" has no place on this express way. Nor need it have, for today JK crystals are available that exceed your most critical stability requirements for interference-free land, sea and air radio communications. Crystals of this extreme stability are also serving as filter resonators for radar, for power line carrier communications and telemetering, and as ultra-precision time bases. And our research goes on. JK Glasline crystals, and JK Ovens, offer "packaging" that combines unprecedented enrivonmental control with the compactness your new equipment requires. So precise, so rugged, so protected are these JK crystals that their performance can now be Certified to previously unheard of stability tolerances. So don't let frequency management problems delay your development programs. JK engineering can be the key needed to unlock tomorrow's door for you.

JK-G9: Precision "Glasline" quartz crystals, sealed in evacuated glass for cleanliness and protection, over a complete range of 800 cycles to 5 mc.

TO COMPLETE THE ENVIRONMENTAL PICTURE: THE NEW JK09 OVEN

To the protection JK Glasline design provides against moisture, contamination, vibration and barometric pressure, the JKO9 oven adds control of temperature. It is production tooled for economy and uniformity, is small and light (1.28" x 1.70" ±1.5 oz.), and is capable of maintaining a temperature constant to within ±*1°C over a range of −55° to +100°C*. Here is an oven that matches the performance of many, massive multi-stage heaters — an example of JK's ultra-stable miniaturization program.





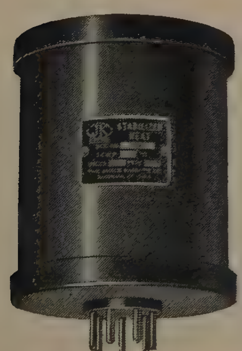
JK-G4: "Glasline" Crystal Filter Resonator. For broad filter applications such as power line carrier communications and telemetering. Frequency range 50 to 200 kc.

PRODUCTS

THE JAMES KNIGHTS COMPANY

Crystals for the Critical - Sandwich, Illinois

JK STANDARD MILITARY AND COMMERCIAL TYPE CRYSTALS

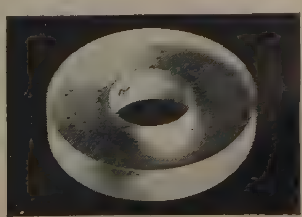


TEMPERATURE CONTROL OVENS: Available for a wide range of applications.

Quality and service are James Knights traditions that apply to the simplest JK crystals as well as to the most complex, and apply to our smallest customer as well as our largest. So whatever your requirements — look to James Knights as a dependable, cooperative source for quality, price and delivery.

PRESSURE MOUNTED: A complete line of commercial and military types.

MILITARY TYPES: Hermetic sealed, metal cased, in frequency ranges from 16 kc to 100 mc.



ULTRA-SONIC TRANSDUCERS: Carefully oriented and processed to your specifications, in a variety of shapes with holes, dimples, soldered-on leads, and backing plates. Can be plated with a variety of metals.

Section Meetings

(Continued from page 100A)

Buffalo-Niagara

"An Airborne Weather Radar," by K. F. Molz, Bendix Radio Div.; December 15, 1954.

"Modern Techniques in X-Ray," by Kevin Kilcoyne and "Distortion and Frequency Response in a Binaural Music System," by David Coddington, both students, University of Buffalo; January 19, 1955.

Cedar Rapids

"Corpuscular Streams from the Sun," by Dr. W. O. Roberts, High Altitude Observatory, Boulder, Colorado; November 17, 1954.

"Dynamotor Design Considerations," by E. I. Winquist, Continental Electric Company; December 15, 1954.

Cincinnati

"A High Powered Amplifier for High-Channel VHF TV Service," by W. F. Goetter, General Electric Company; December 21, 1954.

Connecticut Valley

"Recent Applications of Information Theory," by Louis A. DeRosa, Federal Telecommunications Laboratory; December 16, 1954.

Dallas-Fort Worth

"Matching Opportunity with Need," by Dr. J. W. McRae, Sandia Corp.; November 2, 1954.

Talks on Electronics and Geophysics: "What is the Problem," by Dr. D. H. Clewell, Magnolia Petroleum Company, "Principles of Airborne Magnetometer Instrumentation," by T. W. Swafford, Geotechnical Corp., "Electronic Problems on Seismology," by Dr. J. P. Woods, Atlantic Refining Company; November 23, 1954.

"Measurement of Radio Interference," by Joseph Lorch, Empire Devices Production Corp.; November 29, 1954.

"Electronics Industry and I.R.E. in the Southwest," by Dr. A. W. Straiton, Faculty, University of Texas, and "How to Specify Transformers," by Walt Lehnert, Audio Development Company; December 7, 1954.

"Recent Developments in Long Range Communication Techniques," by A. A. Collins, Collins Radio Company (paper co-authored by M. L. Doelz); December 16, 1954.

Dayton

"Systems Management Concepts," by E. G. Uhl, G. L. Martin Co.; December 2, 1954.

"Hi Fidelity," by Julius Knapp, United Transformer Corp.; January 6, 1955.

Detroit

"Airborne Weather Radar System," by K. F. Molz, Bendix Radar Eng'g. Dept.; December 17, 1954.

El Paso

"Experimental Origins of Lie Detection Equipment," by Dr. R. F. Mager; December 16, 1954.

Evansville-Owensboro

"Technical Aspects of an Integrated Audio System," by D. W. Pugsley, General Electric Company; October 13, 1954.

Field trip to Indiana Bell Telephone Company; January 12, 1955.

Fort Wayne

"The Transmission of Electrical Signals in the Central Nervous System," by Dr. W. S. McCulloch, Mass. Institute of Technology; January 6, 1955.

Hamilton

"Trends in Component Design," by A. Ainlay, Rogers Majestic Electronics Ltd.; December 13, 1954.

(Continued on page 118A)

. . . here's new might for miniatures

Dur-Mica DM15 World's Smallest Mica Capacitor

The First Miniature Dipped Mica Capacitors with Parallel Leads.

(Shown Actual Size)

NOW! Also available . . .

El Menco Dur-Mica DM20
1 to 3900 mmf. at 500vDCw
1 to 5100 mmf. at 300vDCw

El-Menco Dur-Mica

IDEAL FOR PRINTED CIRCUITS. Meets all Humidity, Temperature and Electrical Requirements of MIL-C-5 Specifications. El Menco's Dur-Mica DM15 establishes a *"new dimension"* in capacitor performance with ranges from 1 to 390 mmf. at 500vDCw and 1 to 510 mmf. at 300vDCw. A new, tougher phenolic casing provides temperature co-efficient and stability equal to or better than characteristic F in all but the lowest capacity values . . . efficient operation at temperatures as high as 125°C. El Menco's Dur-Mica DM15 can be used in a variety of transistor circuits and other miniature electronic equipment in military and civilian applications.

Sells for Less than the famous El Menco CM-15 — Provides Economy of Size with Maximum Performance and Widest Application.

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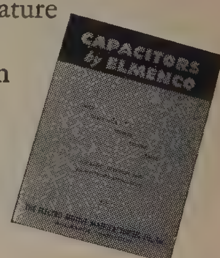

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Section Meetings

(Continued from page 112A)

Hawaii

Demonstration and description of circuitry, International Business Machine Electronic Calculator #604, by J. P. Erdman, IBM Co.; October 13, 1954.

"CAA Airport Surveillance Radar ASR-2," A description of design and circuitry, by Frank Kadi, CAA; November 10, 1954.

Symposium and discussion of long range navigational aids, by Capt. P. V. Colmar, Lt. Cmdr. R. L. Harris and Stanley Pickarski, U. S. Coast Guard; January 12, 1955.

Indianapolis

"Fact, Fancy and Future of Color Television," by R. B. McGregor, Arvin Industries; October 18, 1954.

Inyokern

"Electronic Test Equipment: The Viewpoints of the Manufacturer and User," discussed by W. R. Hewlett and R. E. Rawlins, both of Hewlett-Packard Co.; December 13, 1954.

Israel

"Modern Methods in Electronic Navigation," by J. Halberstein; December 12, 1954.

Ithaca

"Core Memories for High Speed Digital Computers," by W. N. Papian, Mass. Institute of Technology; January 6, 1955.

Little Rock

"Behind the Scenes in Television," by program and production staff of station KARK; December 14, 1954.

Long Island

"Ferromagnetism Applied to Microwave Circuit Techniques," by J. H. Rowen, Bell Telephone Labs.; December 14, 1954.

"Compression-Expansion of Speech," by Dr. W. L. Everitt, University of Illinois; January 11, 1955

Milwaukee

"Photoconductive process," by Dr. Rudolf Frerichs, Faculty, Northwestern University; November 18, 1954.

"Ferrites," by Dr. H. M. Schlicke, Allen Bradley Co.; December 9, 1954.

New York

"Modern Magnetic Ferrites and Engineering Applications," by C. D. Owens, Bell Telephone Labs.; December 1, 1954.

North Carolina-Virginia

"Television Coverage of Coronation of Queen Elizabeth II of England," by Prof. Maurice Valet, Catholic Faculty of Sciences of Lyons, France; December 10, 1954.

"Granpa," by Dr. V. S. Carson, North Carolina State College; January 14, 1955.

Northern New Jersey

"Design Principles in Negative Feedback Transistor Amplifiers," by F. H. Blecher and T. R. Finch (Mr. Finch delivered the paper), both of Bell Telephone Labs.; January 12, 1955.

Oklahoma City

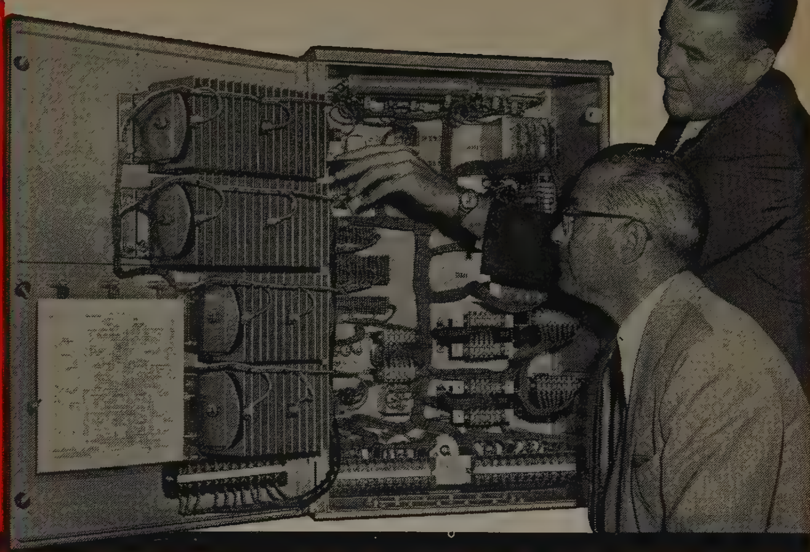
"Computers Without Numbers," by J. B. Wiley, Oklahoma University; December 14, 1954.

"The Bell Solar Battery," by H. J. McMains, Southwestern Bell Telephone Company; January 11, 1955.

(Continued on page 122A)

Costs go tumbling...

when standard Radio Receptor rectifiers do the work of specials in magnetic amplifier applications



Mr. Dornhoefer (upper right) inspects current production of magnetic amplifier regulator with Mr. J. F. Hysler. Rating of the motor generator set it regulates:
Output: 5KVA, 120 V, 3 phase, 400 cycles.
Input: 175 to 345 V, DC.
Regulation accuracy: ±0.5% on both voltage and frequency.
Ambient temperature: 50°C.
Made for and has passed H. I. shock tests.

Here's a magnetic amplifier regulator just off the production line at Regulator Equipment Corp.'s plant in Paterson, N. J. It includes eight RADIO RECEPTOR selenium rectifier types using standard quality cells for a total of 18 units in all, and regulates the voltage and frequency of 400 cycle motor generator sets aboard many of Uncle Sam's submarines.

"The decision to use standard stacks," says Warren Dornhoefer of Regulator Equipment Corp., "is governed by such factors as desired magnetic amplifier performance, reactor core material, ambient temperatures, power supply frequency and many others. Naturally we aim for the right combination to give best overall results."

"It has been our experience," continues Mr. Dornhoefer, "that the standard RADIO RECEPTOR stacks we use perform highly satisfactorily in this mag-amp application and in others we have designed and produced. When we see such excellent results from the regular stacks we prefer to be realistic—particularly when delivery and cost are factors."

Naturally, stock rectifiers are not always the answer for every magnetic amplifier circuit. We can and do supply specials where necessary. We suggest you let us study your specs the next time you require rectifiers for this purpose. Chances are we can save you money—and time!

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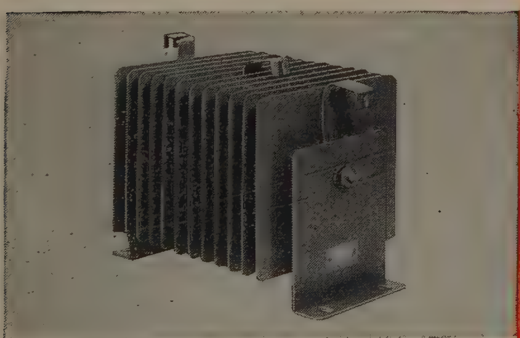

One of the Radio Receptor rectifiers incorporated into Regulator Equipment Corp.'s magnetic amplifier regulator.

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Section Meetings

(Continued from page 118A)

OTTAWA

"Operational Research Applied to Engineering," by Dr. G. R. Lindsey, Defence Research Board; December 9, 1954.

PHILADELPHIA

"Automation in the Automotive Industry," by J. E. Cunningham, Wilson Automation Company; October 14, 1954.

"Automatic Fabrication in the Electronics Industry," by Cledo Brunetti, General Mills; October 21, 1954.

"Etched Circuits," by Donald Mackey, R.C.A., and "Mechanization Techniques and Standards for the Use of Printed Circuitry," by Frank B. Iles and J. J. Graham, both of R.C.A.; October 28, 1954.

"Flexible Control of Electronic Assembly Automation," by G. W. Gamble, G. E. Adv. Elect. Center, and "Modular Approach to Mechanized Assembly," by R. L. Henry, A.C.F. Corp.; November 4, 1954.

"Instrumentation and the Analogue Computer in Process Control," by D. P. Eckman, Case Institute of Technology, and "Important Developments Affecting Automation in Chemical Plants," by V. F. Hanson, E. I. DuPont de Nemours & Co.; November 11, 1954.

"Management Aspects of Automation," by W. W. Beardslee, General Electric Company: November 18, 1954.

"Solar Batteries," by D. C. Reynolds, Wright-Patterson A.F.B., and Gordon Raisbeck, Bell Telephone Labs.; December 13, 1954.

PHOENIX

"Etched Circuits," by Larry Ott and George McLaughlin, both of Hughes Aircraft; December 9, 1954.

PITTSBURGH

"LORAC—Radiolocation by Phase Comparison," by W. R. Hunsicker, Seismograph Service Corp.; December 13, 1954.

"Junction Transistors Used as Switches," by Dr. R. L. Bright, Westinghouse Electric Corp.; January 10, 1955.

PORTLAND

"Electronic Aspects of Civil Defense," by James Deer, Iron Fireman Company; Films, "Main Street USA" and "Use of Radiological Instruments"; December 22, 1954.

Uses of germanium and silicon as transistors and photocells, demonstrated by Dr. W. H. Brattain, Bell Telephone Labs.; January 12, 1955.

PRINCETON

"Electronics at Brookhaven," by R. L. Chase, Instrumentation Div. of Brookhaven; December 9, 1954.

"An Outguessing Machine, A Sophisticated No-Arm Bandit," by Dr. D. W. Hagleberger, Bell Telephone Labs.; January 13, 1955.

ROCHESTER

"Federal Airways Operation and Radio Control of Aircraft," by G. R. Freitag, Rochester Airport; December 7, 1954.

ROME-UTICA

"Nuclear Reactor Design," by G. D. Roy; January 4, 1955.

SACRAMENTO

"Transistors and Their Application in Switching Circuits," by J. M. Pettit, Stanford University; January 14, 1955.

(Continued on page 130A)

Responsibility for the contents of papers published in the PROCEEDINGS OF THE IRE rests upon the authors. Statements made in papers are not binding on the Institute or its members.

PROCEEDINGS OF THE IRE

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CONTENTS

Changes of address (with advance notice of fifteen days) and communications regarding subscriptions and payments should be mailed to the Secretary of the Institute, at 450 Ahnaip St., Menasha, Wisconsin, or 1 East 79 Street, New York 21, N. Y.

John N. Dyer

DIRECTOR, 1955–1956

John N. Dyer was born in Haverhill, Massachusetts on July 14, 1910. Attending Massachusetts Institute of Technology, Mr. Dyer in 1931 received the B.S. degree in electrical engineering.

He was in charge of radio engineering from 1933 to 1935 for the Byrd Antarctic Expedition II. Upon his return to this country, he joined the General Engineering Department of CBS, and in 1937 he became Assistant Chief Television Engineer. At the Radio Research Laboratory of Harvard University, Mr. Dyer in 1942 led the group which developed vhf transmitters. Early in 1944 he became a Director of the American British Laboratory of the National Defense Research Committee in England. It was his responsibility to assist the services in making maximum use of the countermeasure equipment developed at Harvard and other U. S. laboratories. After VE day Mr. Dyer was appointed head of the Radio Research Laboratory's field division. Following this work he joined the Airborne Instruments Laboratory where, in 1950, he became Director of the Research and Engineering Division and, in 1951, Vice-President and a member of the Board of Directors. Under Mr. Dyer's supervision, the former Radar and Air Navigation Section conducted air traffic control studies and experiments in the use of radar at L and S bands for terminal air traffic control. In addition, he conducted general studies on all phases of air navigation and air traffic control for the Air Transport Association. These studies pioneered in the use of ground radar in civil aviation and led to equipment development at the Airborne Instruments Laboratory of rf components, high performance receivers, special antennas, and improvements in MTI techniques. The work of Mr. Dyer's section also involved video mapping, circular polarization, track-while-scan, and taxi radar.

Mr. Dyer joined the IRE in 1930 as a Junior Member. He became an Associate in 1932 and a Senior Member in 1945. In 1949 he received the Fellow Award "for administrative and technical contributions to radio, including polar-expedition communications and important wartime radio countermeasures." Mr. Dyer has served on the Navigation Aids Committee and as a representative for the Long Island Subsection. In 1950 he was elected Chairman of the subsection.

The Greatest Show on Earth

On the evening of Tuesday, March 15, the vanguard of a giant caravan of trucks will begin pouring into New York City from every part of the country. By early Wednesday morning these ten- and fourteen-wheel monsters will be backed up for eight blocks outside the Kingsbridge Armory and the nearby Kingsbridge Palace in the Bronx, waiting to disgorge hundreds of boxes. The boxes will be many sizes and shapes, but their labels will be same: "1955 IRE National Convention."

As the boxes stream into the mammoth Armory, a 1,000-man army of movers, riggers, carpenters, iron workers, painters and electricians will begin the task of unpacking and setting up the exhibits. Their movements will be controlled by a platoon of supervisors and messengers mounted, believe it or not, on bicycles so that they can speedily reach any of the 704 exhibit locations on the enormous four-acre drill floor.

By the end of the day, 100 trailer trucks will have filed to the loading platform, been unloaded and sent on their way. The following morning the line of trucks will still be eight blocks long.

In fact, the unloading process will continue at a rate of 100 trucks a day for four more days before the seemingly endless stream ceases. Finally, on Sunday night, a four-mile line of trucks will be empty, the unpacking of 20,000 boxes will be completed, the exhibits will all be assembled, two lecture halls will be equipped, registration facilities will be set up, and a 1,000,000-watt substation will be installed and connected to the vast array of booths.

Sunday night will see the midnight oil burning also at the Waldorf-Astoria Hotel in mid-town Manhattan, as the work of setting up six more lecture halls, convention headquarters, and sixty feet of registration desks nears completion. What will not be visible to the eye is the seven months of hard work which 175 volunteer committee members spent in planning and making arrangements for the technical program, banquet, cocktail party, facilities, registration, publicity, and so on ad infinitum.

The magnitude, as great as it is, of all these preparations and labors is far overshadowed by the importance of what will take place on March 21–24. On those dates 40,000 people are going to attend the greatest engineering show and convention on earth. They are coming because they cannot afford to miss it.

One of the things they cannot afford to miss is the most extensive program of technical papers in IRE history. A grand total of fifty-five sessions will be presented during four days at the Waldorf-Astoria, Kingsbridge Armory, and Belmont Plaza. This means there will be something like 275 papers and discussions by the leading authorities in the field. They will reveal a host of new technical developments in just about any subject you could mention (e.g., remote control of space stations, designing machines to simulate the behavior of the human brain, high fidelity, to name a few). The complete menu of this savory technical feast begins on page 348 of this issue.

Another reason that 40,000 people are coming is the fabulous Radio Engineering Show. You would have to see it to believe it, because it defies description. Suffice it to say that the visitor who tours the 704 exhibits will learn more about the radio-electronic industry in four hours than he otherwise could in four months, or probably four years.

There is one last and especially important thing that 40,000 people are coming to see—each other. The value which an engineer derives from meeting other engineers and exchanging ideas with them is immeasurable. In two minutes he can learn from the experience of another what it would take months to learn by himself. The convention offers unlimited opportunities of enjoying this stimulating experience of personal contact, not only at the sessions and exhibits, but also at the other features of the convention: the annual meeting of the IRE Monday morning, the cocktail party Monday evening, and annual banquet Wednesday night.

Many rich rewards await you at the IRE National Convention. And they cannot be had by staying home and reading about them.

—The Managing Editor

A Large Signal Theory of Traveling-Wave Amplifiers*

P. K. TIEN†, ASSOCIATE, IRE, L. R. WALKER†, AND V. M. WOLONTIS†

Summary—The large-signal or nonlinear behavior of the traveling-wave amplifier is calculated in this paper by numerically integrating a set of equations of motion, which is essentially that of Nordsieck, but includes the space-charge repulsion between the electrons. The calculations were made assuming a loss-free circuit and a small coupling between the circuit and the beam. A method of computing the space-charge field, by summing the forces between the electrons, is fully described. It is found that in the large signal theory a parameter k is required in addition to the space-charge parameter QC. $1/k$ measures the range of action of the space-charge forces and is proportional to the beam radius. The other parameters including QC are those defined in the linear theory.

Twenty cases covering useful ranges of the design and the operating parameters have been computed. The numerical integrations were done by type 701 IBM equipment. The amplitude and the phase of the circuit wave as functions of the distance are given for all the cases. It is found that the limiting efficiency increases with the electron injection speed, as found by Nordsieck, and increases with the space-charge parameter QC when QC is small, but decreases rapidly for larger QC. The efficiency also increases with the beam radius, or $1/k$, provided the assumption that the field and the current are uniformly distributed over the cross section is valid. Finally the electron motion in high level operation is analyzed. The effect of QC and of k on efficiency are explained. The study indicates several possible ways of improving the limiting efficiency.

Introduction

THE HIGH LEVEL or nonlinear behavior of the traveling-wave amplifier has been analyzed by Nordsieck.[1] Among the simplifications which were made to make the problem tractable was a neglect of the space-charge forces between the electrons. At high level, when electrons start to overtake one another, the space-charge field is generally not negligible, and in some cases may even be much greater than the circuit field. It is therefore the purpose of this paper to extend Nordsieck's model to include the space-charge interaction. For a better understanding of the problem, we shall review briefly some important features of Nordsieck's analysis and then come to the problem of space charge.

Let us consider the major assumptions which are made. In the first place, the model used is one dimensional, in the sense that we shall speak of a charge density, current and velocity which are functions only of distance along the tube (and of time). Thus, all electrons with a given co-ordinate z are supposed to be acted upon by the same rf fields, whereas, in fact, the circuit field will be stronger and the space-charge field will be weaker for the outermost electrons. The "electrons" considered in the model may be considered to represent the average behavior of all the electrons in the beam at a given z. In the same way the excitation of the circuit by the beam is represented by the integrated effect on the circuit of all the charges at a given z. The circuit, furthermore, will be supposed to have a finite impedance for the fundamental angular frequency, ω, only, or stated more specifically, only the fundamental frequency component of the beam current excites a propagating mode on the circuit. This is generally true in actual tubes as the circuit impedance is small for harmonics. It is also assumed that the coupling of the beam to the circuit is weak, or using Pierce's[2] terminology, that the gain parameter, C, is small. This has the effect of insuring that the circuit voltage changes only slightly in one wavelength and that the dc beam velocity, circuit velocity, and the resultant wave velocity are all close together. The latter proviso implies that the backward wave on the circuit remains very far from synchronism with the electrons. Because of this lack of synchronism it is considered legitimate to suppose that the effect of the backward wave upon the electrons may be ignored. One calculates therefore only the excitation of the forward wave. The tube is supposed to be of infinite length, so that what is sought is a "natural" nonlinear solution, periodic in time, which shall be the generalization of the growing wave solution of linear theory.

The features of the model so far described are just those of Nordsieck's treatment. In the present analysis we wish to add to this model a representation of space-charge forces, or more precisely, the electrostatic interaction between the electrons. This may be carried out in a way which is consistent with our earlier assumptions, if we suppose that our "electrons" are uniformly charged discs concentric with the helix but not filling it (the latter for these purposes being thought of as a conducting cylinder). It is a straightforward matter to calculate the force exerted by one such disc on another at a given distance. Integrating over the z-distribution of electrons, we find the space-charge force at any point. To the extent that C is small, we are able to calculate the space-charge force at a point with sufficient accuracy although the electron paths are not yet integrated beyond the point in question. This, of course, is essential if we are to carry out a step-by-step integration.

The Equations

For coherence it is desirable here to recapitulate some important features of the equations which will be used in calculating the rf voltage on the circuit and the electron motions inside the beam, although most of the analysis is to be found in Nordsieck. The circuit voltage

* Original manuscript received by the IRE, October 7, 1954; revised manuscript received, December 6, 1954.

† Bell Telephone Labs., Murray Hill, N. J.

[1] A. Nordsieck, "Theory of the large-signal behavior of traveling-wave amplifiers," PROC. I.R.E., vol. 41, pp. 630–637; May, 1953.

[2] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.

is written as

$$V(z, t) = \frac{Z_0 I_0}{C} A(z) \cos \Phi(z, t) \tag{1}$$

$$\Phi(z, t) = \frac{\omega}{u_0} z - \omega t = \theta(z). \tag{2}$$

Here z is measured along the axis of the helix. The helix is supposed to extend to $z = -\infty$, where *the beam entered unmodulated.* ω is the angular frequency of the fundamental, or $2\pi/\omega$ is the period, and u_0, the average electron speed. $\theta(z)$ is the phase angle between the actual helix wave and a wave in synchronism with the average electron speed. Z_0 is the characteristic impedance of the circuit, I_0, the dc beam current, and C, the gain parameter defined by Pierce.[2] The circuit voltage in (1) must satisfy the equation[1]

$$\frac{\partial^2 V}{\partial t^2} - v_0^2 \frac{\partial^2 V}{\partial z^2} = v_0 Z_0 \frac{\partial^2 \rho_\omega}{\partial t^2}, \tag{3}$$

which describes the excitation of the circuit wave by the electrons and is generally known as the circuit equation in traveling-wave tube theory. v_0 is the velocity of propagation on a cold helix. ρ_ω is the fundamental component of the instantaneous electron linear charge density $\rho(z, t)$.

In addition to the circuit equation (3), we have, for the electron motion, the force equation and the equation of conservation of charge. They are respectively

$$m \frac{d^2 z}{dt^2} = e \frac{\partial V}{\partial z} - eE_s \tag{4}$$

and

$$\rho(z, t) \left| dz \right| = \rho(z, 0) \left| dz_0 \right|, \tag{5}$$

where e/m is the ratio of the electron charge to the mass, a positive quantity. E_s is the space-charge field acting on the electrons, which will be discussed in detail later.

Let the z_0's be the initial positions of the electrons at $t = 0$. We shall denote the initial phase of an electron by

$$\Phi_0 = \frac{\omega z_0}{u_0}. \tag{6}$$

So far we have acted as if $\rho(z, t)$ were a continuous function of z and t. As described in the previous section, for the present model we shall replace the continuous electron beam by a set of "electron-discs." Thus, instead of considering the electron beam as a fluid with a given velocity and density as functions of z and t, we shall follow a typical set of charged discs and calculate their velocities and positions. This is necessary in large-signal calculations when electrons overtake one another. With this in mind, using (3), (4), and (5), we obtain the following working equations after considerable algebra:

$$\frac{dA(y)}{dy} = -\frac{1}{2\pi} \int_0^{2\pi} \sin \Phi(\Phi_0, y) d\Phi_0 \tag{7}$$

$$\frac{d\theta}{dy} + b = \frac{1}{2\pi A(y)} \int_0^{2\pi} \cos \Phi(\Phi_0, y) d\Phi_0 \tag{8}$$

$$2 \frac{\partial q(\Phi_0, y)}{\partial y} = 2A(y) \sin \Phi(\Phi_0, y) - \frac{e}{u_0 m \omega C^2} E_s \tag{9}$$

$$\frac{\partial \Phi(\Phi_0, y)}{\partial y} = -\frac{d\theta(y)}{dy} + 2q(\Phi_0, y). \tag{10}$$

Here (7), (8), (9), and (10), except for the space-charge term E_s, are Nordsieck's equations[1] (13), (14), (17), and (20). z is replaced by:

$$y = C \frac{\omega}{u_0} z. \tag{11}$$

This means that distances are measured in units of $\lambda_e/2\pi C$, where λ_e is the wavelength along the beam. The velocity at y, of an electron of an initial phase Φ_0 is expressed as

$$u(\Phi_0, y) = u_0 + 2Cu_0 q(\Phi_0, y). \tag{12}$$

Here u_0 and $2Cu_0q$ are respectively dc and ac velocities. $\Phi(\Phi_0, y)$ is the phase of the circuit voltage which an electron with an initial phase Φ_0 sees at y. In studying the electron motions, $\Phi(\Phi_0, y)$'s are the time-phases of the electrons, or more precisely, these values give a distribution of the electrons in time at the position y. In a small interval of y, a distribution in ωt at y may also be considered as the distribution in $\omega z/u_0$ at t. It is obvious that electrons whose initial phases are the same or differ by an integral multiple of 2π experience the same circuit and space-charge fields and should have the same motion. It is, therefore, only necessary for us to study the motions of the electrons whose initial phases cover an interval of 2π. b and C, as defined by Pierce, are respectively

$$b \equiv \frac{u_0 - v_0}{Cu_0}$$

$$C^3 \equiv Z_0 \frac{I_0}{4V_0}.$$

Here V_0 and I_0 are respectively the beam voltage and beam current, I_0 being considered as a positive quantity.

We can easily give a physical interpretation of these equations. Eq. (7) indicates that the rate of increase of the circuit voltage depends upon the phases of the circuit voltage at the positions of the electrons. This is equivalent to saying that the increase of the circuit voltage is mainly attributed to the density modulation of the beam. When C is small, as we have assumed previously, the rf velocity is small compared to the dc velocity. The convection current in the beam is therefore approximately equal to the product of the electron linear charge density and the dc velocity, which in our analysis is proportional to $-d\Phi_0$. On the other hand, the electric field of the circuit is $-A(y) \sin \Phi$. The negative of the product of the electric field and the convec-

tion current, which is proportional to $-A(y)\sin\Phi\, d\Phi_0$, integrated over all the electron-discs, is the total power contributed by the beam to the circuit. It should be proportional to $[dA^2(y)/dy]$, which is equal to $2A(y)[dA(y)/dy]$. Cancelling $A(y)$ in $\int A(y)\sin\Phi\, d\Phi_0$ and $2A(y)[dA(y)/dy]$, and equating them through a proper proportioning constant, we obtain exactly the equation (7). In the small-signal theory for small C, the ratio of the rf to dc velocity is a quantity much smaller than that of the rf to the dc linear charge density. It is therefore appropriate in the small-signal theory for small C to neglect the rf velocity in comparison with the dc velocity. The same simplification in the large-signal theory limits the value of C for which our theory applies. In the later calculations, the value of $2q$ near the saturation level is found to be about 3. If, for example, $C=0.02$, the error involved will be only 6 per cent. At $C=0.1$ however, the error amounts to 30 per cent.

To understand (7) and (8) better, we shall compare them with the equation derived by Pierce for the field in the circuit caused by an impressed current.[3] Pierce's equation is

$$V = \frac{-\Gamma\Gamma_1 K}{(\Gamma^2 - \Gamma_1^2)}\, i. \tag{13}$$

In his notation, Γ and Γ_1 are the propagation constants of the actual and the cold helix waves respectively. i is the rf beam current, and K, the circuit impedance. In the small-signal theory, the circuit wave varies as $e^{(\mu_1+j\mu_2)y}$, where μ_1 and μ_2 are respectively Pierce's[2] x_1 and y_1, and $j=\sqrt{-1}$. $[dA(y)/dy]$ is then proportional to $\mu_1 e^{\mu_1 y}$, and

$$A(y)\left[\frac{d\theta(y)}{dy} + b\right],$$

proportional to $(\mu_2+b)e^{\mu_1 y}$. In a later section, we shall indicate that

$$-\frac{1}{2\pi}\int_0^{2\pi} \sin\Phi d\Phi_0,$$

in (7) and

$$-\frac{1}{2\pi}\int_0^{2\pi} \cos\Phi d\Phi_0$$

in (8), are proportional, respectively to the parts of the beam current, which are in-phase (or 180 degrees out-of-phase) and 90 degrees out-of-phase with the axially directed circuit electric field. Eqs. (7) and (8) thus express a relation between the rf beam current and the circuit electric field, and show that in the small-signal case, the current, like the circuit field, has an amplitude varying as $e^{\mu_1 y}$ but the current leads the field in time phase by $\tan^{-1}(\mu_2+b)/-\mu_1$. This relation is identical with (13) as expected.

[3] Pierce, *op. cit.*, eq. (2.10).

Eq. (9) is merely a transform of the force equation, stating that the acceleration of an electron is proportional to the sum of the circuit and the space-charge fields. Eq. (10) means that the phase shift in circuit voltage experienced by an electron in an interval of dy depends upon two factors, the phase delay of the circuit voltage in that interval, $d\theta/dy$, and the electron rf velocity, $2q$, as it should.

A study of these equations indicates that we may keep certain electrons in a favorable position to contribute power to the circuit by minimizing the quantity $d\Phi/dy$ in (10) for these electrons. This requires $(d\theta/dy) \cong 2q(\Phi_0, y)$ which can be achieved by properly adjusting the parameter b in (8) as a function of y. It is therefore possible to increase the efficiency of the device by properly varying the helix pitch or dc electron speed along the tube. The limitation comes, however, when the electrons are so much decelerated, that the amplitude of the rf velocity exceeds the dc value. In order to contribute power, electrons unfortunately must be in a retarding field.

The Space Charge

We now want to introduce space charge or electrostatic interaction into the problem in a manner consistent with other features of the model. It is clear that in a one-dimensional hydrodynamic picture, the charge $\rho(z', t)dz'$ lying in z' to $z'+dz'$ at a time t, should produce, at a point z and at a time t, a field

$$-B(z'-z)\rho(z', t)dz', \tag{13}$$

where $B(z)$ is an appropriate function which will be discussed in detail later. The total field at z would then be

$$E_s = -\int_{-\infty}^{+\infty} B(z'-z)\rho(z', t)dz'.$$

From (5) we have

$$\rho(z', t)\,|\,dz'\,| = \rho(z_0, 0)\,|\,dz_0\,| = \frac{-I_0}{u_0}\,|\,dz_0\,|.$$

Thus,

$$E_s = \frac{I_0}{u_0}\int_{-\infty}^{+\infty} B[z(z_0', t) - z(z_0, t)]\,|\,dz_0\,|. \tag{14}$$

Here z_0' and z_0 are the initial positions of the electrons which at time t and at z' and z respectively. The integral in (14) may be replaced, of course, by a sum over individual electrons.

Two questions arise immediately. The first is whether or not the fact that the space-charge field at a point depends upon the disposition of charges ahead of the point in question makes it impossible to proceed with a step-by-step integration of the equations of motion. The second question is that of a sensible choice for the function $B(z)$.

Consider the first problem. $B(z)$ is obviously a rapidly decreasing function of $|z'-z|$. As a matter of fact, we will later approximate this function by an exponential function $e^{-k(\omega/u_0)|z'-z|}$, where the value of k is generally between 1 and 5. Taking $k=2$ for example, the function $e^{-k(\omega/u_0)|z'-z|}$ decreases to 0.00185 of its maximum value for $(\omega/u_0)|z'-z| = \pi$. It therefore seems safe for us to compute the space-charge field at z by considering only the electrons in an interval between $(\omega/u_0)z+\pi$ and $(\omega/u_0)z-\pi$. For small C, $\theta(y)$, and $A(y)$ only vary slightly in a few electronic wavelengths. We may thus substitute for $z(z_0', t)-z(z_0, t)$ the expression

$$\frac{u_0}{\omega}\left\{\Phi[\Phi_0', y(\Phi_0', t)] - \Phi[\Phi_0, y(\Phi_0, t)]\right\}, \quad (15)$$

where $(u_0/\omega)\Phi[\Phi_0', y(\Phi_0', t)]$ means evidently a distribution of the electrons in z at a time t. A distribution of the electrons in z at constant t, in the vicinity of the point of interest, is approximately the same as the distribution in t (expressed in the unit of z) at constant z, provided that the circuit wave propagates with a constant amplitude at a constant speed and that the ac velocities of the electrons are small compared with the dc velocity. That is, to the extent that $\theta(y)$ and $A(y)$ are considered constant in an interval in which the values of $B(z'-z)$ cannot be ignored, (15) to the accuracy of order C, may be further replaced by

$$\frac{u_0}{\omega}\left\{\Phi[\Phi_0', y] - \Phi[\Phi_0, y]\right\}, \quad (16)$$

where

$$y = y(\Phi_0, t).$$

We thus obtain the space charge field at y,

$$E_s = \frac{I_0}{\omega}\int_{-\infty}^{+\infty} B\left\{\frac{u_0}{\omega}\left[\Phi(\Phi_0+\phi, y) - \Phi(\Phi_0, y)\right]\right\} d\phi. \quad (17)$$

Here Φ_0 is the initial phase of the electron which, at the time t, is situated at y, where the space-charge field E_s is evaluated. Thus, for an electron of an initial phase Φ_0, the term E_s is calculable at y from a knowledge of Φ's for all the other electrons when they are at y. Since $\Phi[\Phi_0+\phi, y]$ is periodic in ϕ, this means knowing what happens to the electrons in one cycle of ϕ at y.

Turning now to the problem of how to choose $B(z)$, it is clear that the recipe most consistent with our model is to calculate the force between two infinitely thin discs of charge with a separation, z. To do this, we replace the helix by a conducting cylinder of equal radius and assume that the charges are uniformly distributed over the discs. We can calculate, by Green's function for the cylinder,[4] the potential for one disc, and thus find the force on the second. This yields for $B(z)$

$$B(z) = \frac{1}{2}\,\frac{1}{\epsilon_0\pi r_0^2}\sum_n e^{-(\gamma_n(r_0/a))z/r_0} \cdot \left[\frac{2}{\gamma_n}\,\frac{J_1\left(\gamma_n \frac{r_0}{a}\right)}{J_1(\gamma_n)}\right]^2 \operatorname{sgn} z, \quad (18)$$

where

$$\operatorname{sgn} z = 1 \quad \text{for} \quad z' - z > 0$$
$$= -1 \quad \text{for} \quad z' - z < 0$$

and

$$J_0(\gamma_n) = 0,$$

where the J's are Bessel functions. In (18) r_0 and a are the radii of the disc and the helix respectively. $2\pi r_0^2\epsilon_0 B(z)$ is plotted in Fig. 1 as a function of z/r_0 for

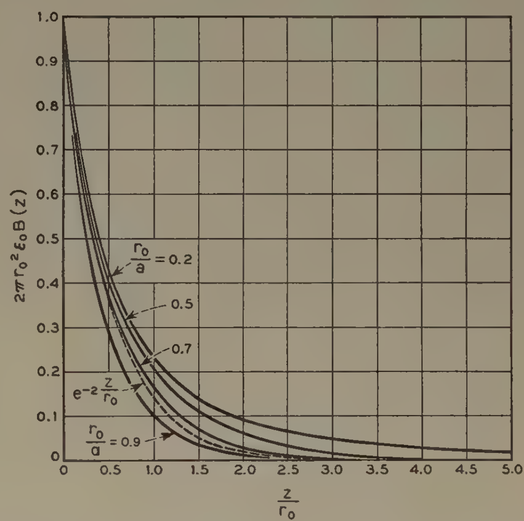


Fig. 1—The solid curves show $2\pi r_0^2\epsilon_0 B(z)$ versus z/r_0, calculated for the force between two electron-discs inside a conducting cylinder, and the dotted curve indicates that computed from the universal function e^{-2z/r_0}.

several values of r_0/a. When z/r_0 is small, the curves lie close together and are closely approximated by exponentials of the form $e^{-\alpha z/r_0}$. Where α is a constant in Fig. 1 the function

$$2\pi r_0^2\epsilon_0 B(z) = e^{-2z/r_0}\operatorname{sgn} z \quad (19)$$

is shown as a dotted line for comparison. The deviations from exponential form occur for large values of z/r_0, where the values of $B(z)$ are small and the space-charge term in the force is relatively unimportant. Because of the convenience of an exponential function in the space-charge computations, it is desirable to use this form of approximation to $B(z)$. If z be replaced by (16), the space-charge term in (9) may be expressed as

$$\frac{e}{m\omega C^2 u_0} E_s = g\int_{-\infty}^{+\infty} e^{-k\left|\Phi(\Phi_0+\phi,y)-\Phi(\Phi_0,y)\right|} \cdot \operatorname{sgn}\left\{\Phi(\Phi_0+\phi, y) - \Phi(\Phi_0, y)\right\} d\phi, \quad (20)$$

[4] W. R. Smythe, "Static and Dynamic Electricity," McGraw-Hill Book Co., Inc., New York, N. Y.; 1939.

with

$$g = \frac{eI_0}{2m\omega^2C^2u_0\epsilon\pi r_0^2} = \frac{1}{2}\frac{\omega_p^2}{\omega^2C^2} \quad (21)$$

and

$$k = \frac{\alpha}{\frac{\omega}{u_0} r_0}.$$

Here ω_p is the familiar electron plasma angular frequency for a beam of infinite extent. Pierce has emphasized that *Fig. 1 shows α to vary very little with r_0/a and suggests that for simplicity it may be taken equal to 2 as in (19)*. When this choice is made, it may be seen that if $(\omega/u_0)r_0$ runs from 0.4 to 2, as it might in practice, k varies from 5 to 1, as mentioned. $1/2\pi k$, clearly, measures range of force between disc charges in units of beam wavelength. With α fixed, range is proportional to r_0, as might be expected, and independent of helix radius.

In the disc model, then, the space-charge force is characterized by the two parameters, g and k; g measuring the intensity and $1/k$ the range of the force. It is essential to make a connection between these parameters and the quantity QC, which appears in the usual linear theory with space charge. This is effected by making a linearized or small signal theory of the present set of equations (carried out in Appendix I), and we find

$$QC = \frac{g}{2(k^2+1)} = \frac{\omega_p^2}{4\omega^2C^2(k^2+1)}, \quad (22)$$

or

$$4QC^3 = \left(\frac{\omega_p}{\omega\sqrt{k^2+1}}\right)^2. \quad (23)$$

The factor $[1/(1+k^2)]^{1/2}$ in (23) may be considered as a reduction factor for ω_p for a finite beam inside a helix. Denoting this reduction factor by R, and using the formula $2u_0/\omega r_0$ for k, we have

$$R = \left[1 + \left(\frac{2u_0}{\omega r_0}\right)^2\right]^{-1/2}. \quad (24)$$

Pierce[5] has computed values of QC by replacing the helix by a conducting cylinder as we have done and his values agree reasonably with those of Fletcher's[6] more elaborate calculations. We may make a check on our method of evaluating space charge by comparing R calculated from the work of Fletcher and of Pierce with that found from (24). Fig. 2(a) and 2(b) shows such a comparison to be satisfactory.

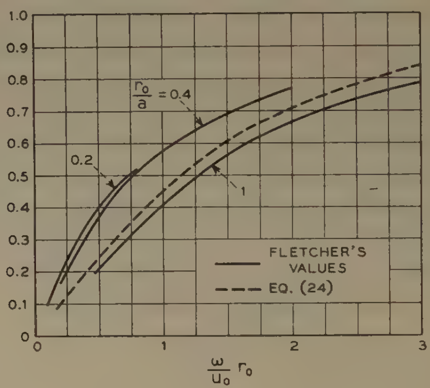
The Space-Charge Parameters QC, g and k

Let us study in more detail the quantities QC, g and k. In (23) and (21), we have respectively

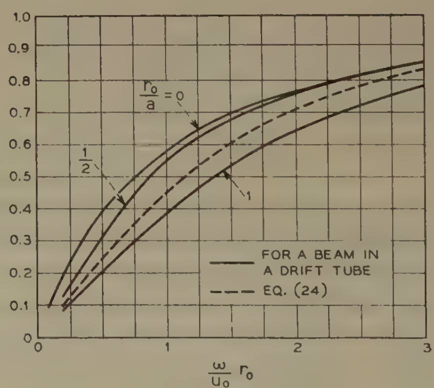
$$QC = \left(\frac{\omega_p R}{2\omega C}\right)^2$$

$$g = \frac{1}{2}\left(\frac{\omega_p}{\omega C}\right)^2.$$

It may be seen that in the QC expression, $\omega_p R$ instead of ω_p is used. ω_p^2 is proportional to the dc electron charge density of the beam. R is the reduction factor of the electron plasma frequency for a beam inside the helix, and can be calculated if the frequency and the radii of the beam and the helix are known. It has a different value for the fundamental and for the harmonic, and is also different for different harmonics. Since the parameter QC is first introduced in the small-signal theory in which the beam current is mainly of the fundamental component, the reduction factor, R, must there be specifically for the fundamental frequency.



(a)



(b)

Fig. 2—The solid curves show the reduction factor, R, of the electron plasma frequency evaluated (a) from the Fletcher's values, (b) for an electron beam inside a conducting cylinder, and the dotted curves indicate that computed by the universal function ϵ^{-22/r_0}.

In the large-signal theory, however, the beam current may contain a considerable amount of harmonic components. Although we may calculate the space charge by evaluating the reduction factors and, thus, the QC parameters for all the harmonic components of the beam, it seems simpler to set up a model consisting of a set of electron discs and compute the space-charge field by evaluating the forces between them in such a way

[5] Pierce, *op. cit.*, pp. 236–238.
[6] R. C. Fletcher, "Helix parameters in traveling-wave tube theory," Proc. I.R.E., vol. 38, pp. 413–417; April, 1950.

that only two parameters, g and k, as shown by (20), actually enter the calculations. The calculation is independent of the helix radius, because we use the approximation

$$k = \frac{2}{(\omega/u_0)r_0}.$$

In the large-signal theory, therefore, we need two parameters such as g and k, whereas in the linearized theory a single parameter QC is sufficient to specify the space charge. Instead of using g and k as the basic space-charge parameters, we often use QC and k for convenience. g, of course, is calculable from QC and k by (22). It is interesting to point out that beams having the same value of QC but different values of k must behave identically in traveling-wave tubes, when the signal is small, since QC is the only space-charge parameter involved in the small signal theory.

Numerical Calculations

The equations used for numerical calculations are (7) to (10), and (20). For most of the cases computed below, the number of points used in evaluating the space-charge integral (20) by the trapezoidal rule is 24, or, stated in physical language, a model based on 24 electron discs in an electronic wavelength is used. The forward integrations, which will be discussed more fully below, are performed in steps of $\Delta y = 0.2$. To determine whether these numbers are adequately chosen, several cases were recalculated with 48 or 96 electron discs and some with $\Delta y = 0.1$. For small values of QC (See Table I, on the next page) the agreement was excellent, and well beyond what could be physically significant. For $QC = 1.0$, and to some extent for $QC = 0.5$, however, substantial variations occurred, resulting in differences of as much as 10 per cent in $A(y)$ for large y. The cause of these variations is the discontinuous nature of the space-charge force, combined with the very large magnitude of the contributions to this force from individual electrons. If the values of Φ for two electrons are nearly equal, a very small change in them may reverse the sign of their difference and, hence, the sign of the space-charge contribution. This phenomenon is also the cause of the minor irregularities appearing in the functions Φ and q past the point of overtaking even for small QC, and it is important to realize that it should not be interpreted as a truncation error in the ordinary sense resulting from inadequate numerical methods. The effect of this phenomenon decreases very slowly with an increasing number of electrons, whereas the calculation time increases steeply. The results presented in this investigation are those calculated, with 48-electron model for cases (13), (14), (18), and (19) (these cases have $QC = 0.5$ or 1.0 and saturate at relatively larger values of y because of large values of b), and with 24-electron model for the rest of the cases. The choice of $\Delta y = 0.2$ was maintained for all the cases.

The numerical calculations were carried out using the IBM Electronic Data Processing Machines Type 701 and Associated Equipment, briefly known as the "701," at IBM World Headquarters in New York City. The high-speed random-access storage of this machine consists of cathode ray tubes capable of holding 4,096 numbers consisting of 35 binary digits and sign—so-called full words, roughly equivalent to 10 digit decimal numbers—or 8,192 numbers or instructions consisting of 17 binary digits and sign, called half-words. Practically unlimited auxiliary storage is available on magnetic tapes and drums, but no use of this equipment was necessary in the present problem. The time required to add two full or half-words in the 701, one placed in the arithmetic unit and the other located anywhere in high-speed storage, is 60 microseconds, and to multiply them, 456 microseconds or less, depending on the operations that follow. Most logical operations, such as deciding between two courses of action depending on the value or sign of the quantity standing in the accumulator, take place in 48 microseconds.[7] At these speeds, the time required for printing out all relevant variables of the problem at the end of each step in the forward integration would be substantially higher than the calculating time. In view of this, the program was designed to print normally only at intervals of 0.4 in y. Furthermore, instead of performing all calculations in so-called floating point form, i.e. with all numbers represented as $a \cdot 10^b$ with a between 1 and 10, the very short calculating time on 701 was achieved by using fixed-point form for all numbers except exponentials occurring in space-charge calculation. Running time per case, for y ranging from 0 to 10, is approximately 10 minutes. The program was prepared by Miss D. C. Leagus.

In view of the nature of the investigation at hand, simplicity rather than high accuracy was used as a guide in the choice of numerical methods. Hence, the forward integration of A, θ, Φ and $2q$ was performed by a central difference formula, e.g.

$$A(y + \Delta y) = A(y) + A'(y + \tfrac{1}{2}\Delta y) \cdot \Delta y.$$

The integrals in (7), (8) and (20) are evaluated by the trapezoidal rule.

The initial conditions were computed from a second-order small-signal theory, which is discussed in detail in Appendix II. The starting point for the forward integration was chosen such that the maximum value of $2q(\Phi, y)$ is approximately 5 per cent of the value of $d\theta/dy$, and is denoted by $y = 0$. The lengthy space charge calculation was performed only at intervals of 0.2. Quantities, such as $d\theta/dy$, demanded by the central difference formula at intermediate points, were obtained by third-order polynomial extrapolation. Subsequently, as a check on machine performance, the value of $d\theta/dy$ thus obtained was compared to the value obtained by applying (8) to extrapolated values of Φ.

[7] For a detailed discussion of this machine, see several articles in Proc. I.R.E., "Computer Issue," vol. 40; October, 1953.

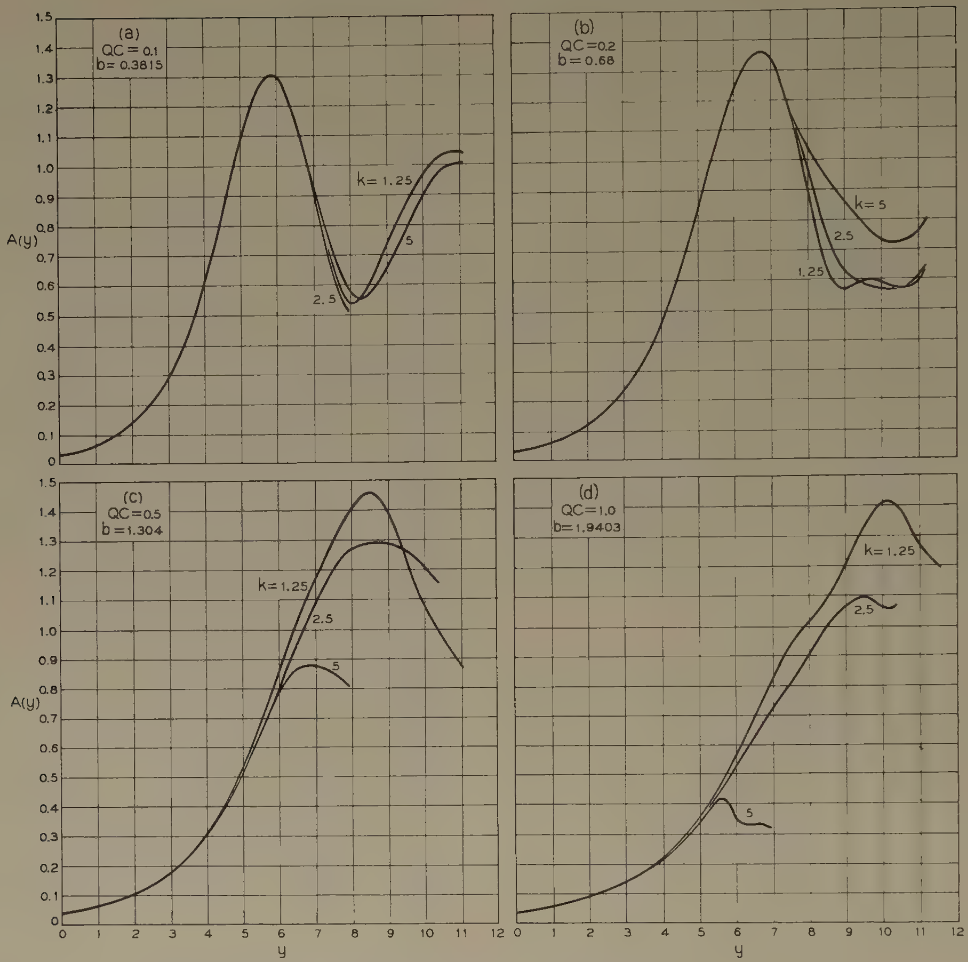


Fig. 3—$A(y)$ versus y curves for (a) cases (1), (2), and (5), (b) cases (6), (7), and (10), (c) cases (11), (12), and (15), (d) cases (16), (17), and (20).

To investigate high level operation for tubes of different signs and with different operating conditions, twenty cases with QC ranging from 0.1 to 1.0, k from 1.25 to 5.00, and b from 0.38 to 2.60, have been computed and are listed in Table I below.

TABLE I

Case Number	QC	k	b	
(1)	0.1	1.25	0.3815	($\mu_1=0.7669$, $\mu_2=-0.7150$)
(2)	0.1	2.50	0.3815	($\mu_1=0.7669$, $\mu_2=-0.7150$)
(3)	0.1	2.50	0.9595	($\mu_1=0.7218$, $\mu_2=-0.9596$)
(4)	0.1	2.50	1.6295	($\mu_1=0.5106$, $\mu_2=-1.261$)
(5)	0.1	5.00	0.3815	($\mu_1=0.7669$, $\mu_2=-0.7150$)
(6)	0.2	1.25	0.6800	($\mu_1=0.6974$, $\mu_2=-0.9196$)
(7)	0.2	2.50	0.6800	($\mu_1=0.6974$, $\mu_2=-0.9196$)
(8)	0.2	2.50	1.200	($\mu_2=0.6564$, $\mu_2=-1.1552$)
(9)	0.2	2.50	1.800	($\mu_1=0.4643$, $\mu_2=-1.4358$)
(10)	0.2	5.00	0.6800	($\mu_1=0.6974$, $\mu_2=-0.9196$)
(11)	0.5	1.25	1.3040	($\mu_1=0.5824$, $\mu_2=-1.4189$)
(12)	0.5	2.50	1.3040	($\mu_1=0.5824$, $\mu_2=-1.4189$)
(13)	0.5	2.50	1.7107	($\mu_1=0.5481$, $\mu_2=-1.615$)
(14)	0.5	2.50	2.1982	($\mu_1=0.3877$, $\mu_2=-1.852$)
(15)	0.5	5.00	1.3040	($\mu_1=0.5824$, $\mu_2=-1.4189$)
(16)	1.0	1.25	1.9403	($\mu_1=0.4962$, $\mu_2=-2.0009$)
(17)	1.0	2.50	1.9403	($\mu_1=0.4962$, $\mu_2=-2.0009$)
(18)	1.0	2.50	2.2800	($\mu_1=0.4670$, $\mu_2=-2.168$)
(19)	1.0	2.50	2.6900	($\mu_1=0.3303$, $\mu_2=-2.371$)
(20)	1.0	5.00	1.9403	($\mu_1=0.4962$, $\mu_2=-2.0009$)

In this table, μ_1 and μ_2, which are respectively Pierce's x_1 and y_1, are also listed for reference. In the small-signal theory, if QC is known, the values of μ_1 and μ_2 can be calculated for any b. It seems wise here to choose b to give particular values of μ_1. Thus, in cases (3), (8), (13), and (18), and in cases (4), (9), (14), and (19), μ_1's are respectively equal to 0.9412 and 0.6658 of their maximum values. For the rest of the cases μ_1's are maximum, that is, at the maximum small-signal gain.

Amplitude and Phase of the Circuit Wave

Among results obtained in calculations, the most valuable probably are the amplitude and phase of circuit voltage as functions of y (Figs. 3–6, on this and the following pages). $A(y)$ is plotted in Figs. 3 and 5 for all cases computed. In Fig. 3, curves are compared for cases of different values of k, in Fig. 5, for cases of different values of b. $\theta(y)$ and $\theta(y)-\mu_2 y$ are similarly plotted in Figs. 4 and 6, in which $\mu_2 y$ is also plotted for comparison.

It is found in the $A(y)$ curves that the circuit voltage first increases with y, and then decreases after reaching a saturation level. The power in the circuit at the saturation level is generally known as the "limiting power."

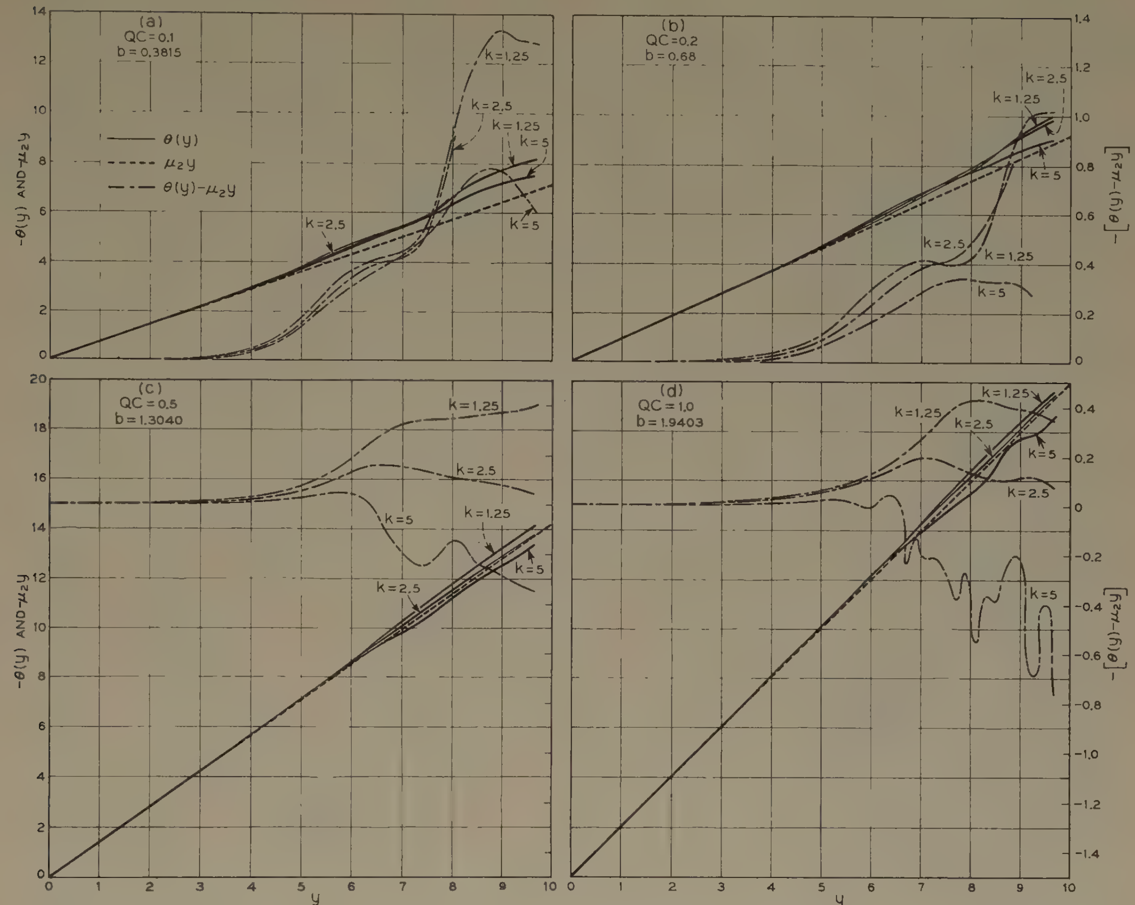


Fig. 4—$\theta(y)$, $\mu_2 y$, and $\theta(y)-\mu_2 y$ versus y curves for (a) cases (1), (2), and (5), (b) cases (6), (7), and (10), (c) cases (11), (12), and (15), (d) cases (16), (17), and (20).

$A(y)$ usually increases again after the first peak and forms another peak and so on. For example, in Fig. 5(b), the $A(y)$ curves for cases (7) and (8) contain three peaks in the range up to $y=20$. For most of the cases, the first peak is the highest one. For $QC=1.0$, and to some extent for $QC=0.5$, the later peaks may have larger, but not much larger amplitudes. $\theta(y)$ does not differ widely from $\mu_2 y$ for almost all the cases.

We have mentioned heretofore that the cases having a common value of QC but different values of k must behave identically at low level. This is so found in Figs. 3 and 4; also that the effect of k is not prominent for small values of QC. This should be so, as k is introduced in the calculation only because of the space charge.

Power and Efficiency

The crcuiit voltage $V(z, t)$ may be calculated from $A(y)$ by (1). The power in the circuit is obviously the time average of V^2/Z_0 that is,

$$\left(\frac{V^2}{Z_0}\right)_{\text{aver.}} = 2CA^2 I_0 V_0. \tag{25}$$

The efficiency of the traveling-wave tube is therefore

$$\text{Eff.} = \frac{2CA^2 I_0 V_0}{I_0 V_0} = 2CA^2 \tag{26a}$$

or

$$\frac{\text{Eff.}}{C} = 2A^2. \tag{26b}$$

The quantity (Eff./C), at the saturation level, or (Eff./C) at the first peak of the $A(y)$ curve [denoted as (Eff./$C)_{\max.}$], is plotted vs QC in Fig. 7(a), on page 269, using k as parameter; in 7(b) using fraction of $\mu_{1(\max.)}$ as parameter, to show how (Eff./$C)_{\max.}$ varies with QC, k and b. Values for $QC=0$ are Nordsieck's. It is seen that (Eff./$C)_{\max.}$ increases with b, a result which has been found by Nordsieck in his calculations without space charge. It also increases with QC for small values of QC, but is reduced sharply at larger values of QC and k. It is also noticed that (Eff./$C)_{\max.}$ decreases as k increases, particularly when QC is large. It should be recalled that k is inversely proportional to the beam radius. In the high power tubes in which QC is generally high, a design with larger beam diameter and yet with a value of $(\omega/u_0)r_0$, such that the field is fairly uniformly distributed over the cross section, is preferred for high efficiency. This has been pointed out by C. C. Cutler as a result of his measurements of overload characteristics.

The same quantity, (Eff./$C)_{\max.}$, is plotted in Fig. 7(c) versus g for all the cases which have the values of b corresponding to the maximum small signal gain. All

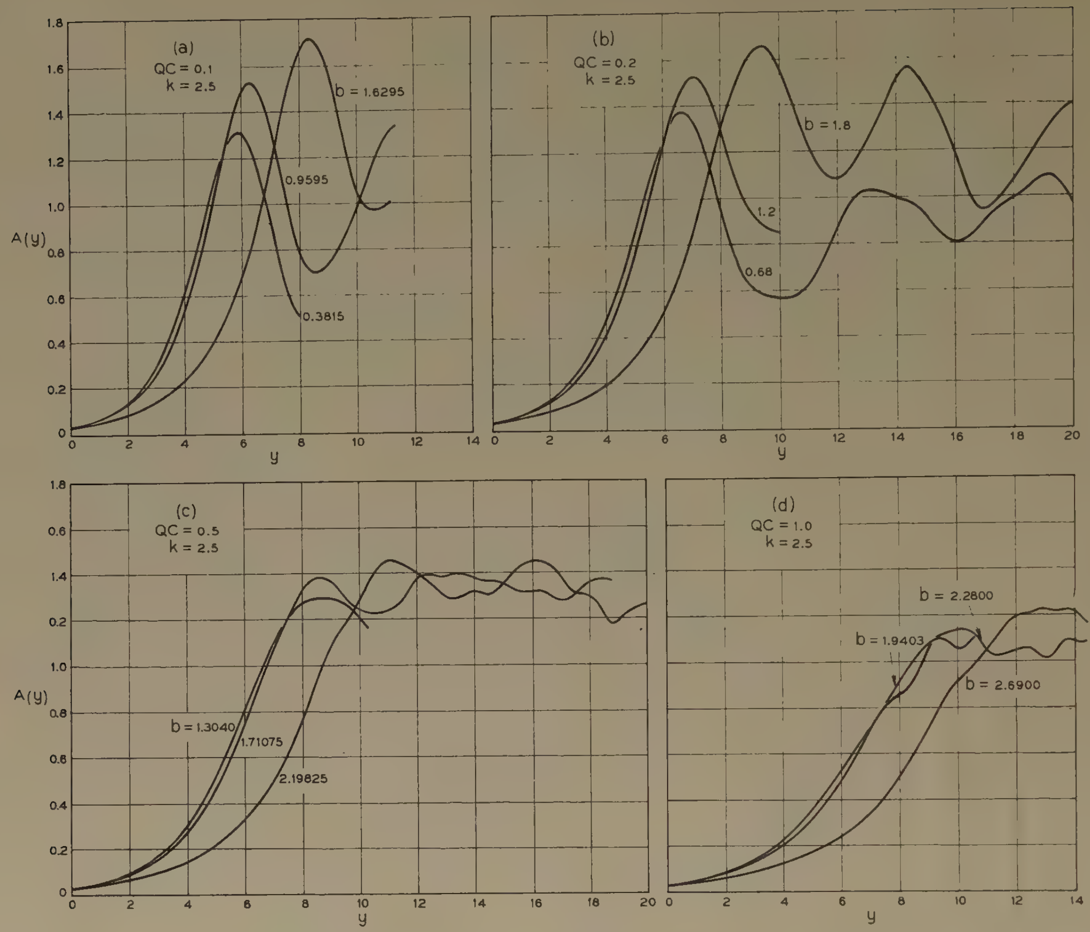


Fig. 5—$A(y)$ versus y curves for (a) cases (2), (3), and (4), (b) cases (7), (8), and (9), (c) cases (12), (13), and (15), (d) cases (17), (18), and (19).

the points are seen to be distributed inside a narrow region whereas they are widely spread in Fig. 7(a). Efficiency varies more with g, which is $\frac{1}{2}(\omega_p/\omega C)^2$, rather than with QC, which is $(\omega_p R/2\omega C)^2$; or with g, the intensity, rather than $1/k$, the range.

The Fundamental Component of the Beam Current

In this section we shall discuss (7) and (8) further. From (5) and for small C, the electron current density of a beam is

$$-I_0 \frac{|d\Phi_0|}{|d\Phi|}. \tag{27}$$

The fundamental component of the beam current is then,

$$i_1(z, t) = -(\sin\Phi)\frac{1}{\pi}\int_0^{2\pi} I_0 \frac{d\Phi_0}{d\Phi'}(\sin\Phi')d\Phi' - (\cos\Phi)\frac{1}{\pi}\int_0^{2\pi} I_0 \frac{d\Phi_0}{d\Phi'}(\cos\Phi')d\Phi'$$

$$= -\frac{I_0}{\pi}\left(\sin\Phi\int_0^{2\pi}\sin\Phi d\Phi_0 + \cos\Phi\int_0^{2\pi}\cos\Phi d\Phi_0\right). \tag{28}$$

In (28), $(1/\pi)\int \sin\Phi d\Phi_0$ is a quantity proportional to the part of $i_1(z, t)$ in-phase or 180 degrees out-of-phase with the circuit electric field, and similarly $(1/\pi)\int \cos\Phi d\Phi_0$ is proportional to the part of $i_1(z, t)$, 90 degrees out-of-phase with the circuit electric field. It should be recalled that the circuit electric field is proportional to $-A(y)\sin\Phi$. Eqs. (7) and (8) therefore mean that the in-phase (or 180 degrees out-of-phase) part and the 90 degrees out-of-phase part of the fundamental component current are proportional to

$$-\frac{dA(y)}{dy} \quad \text{and} \quad \left(\frac{d\theta(y)}{dy} + b\right)A(y)$$

respectively. Thus $i_1(z, t)$ is calculated from values of

$$A(y), \quad \frac{dA(y)}{dy}, \quad \frac{d\theta(y)}{dy} \quad \text{and} \quad b.$$

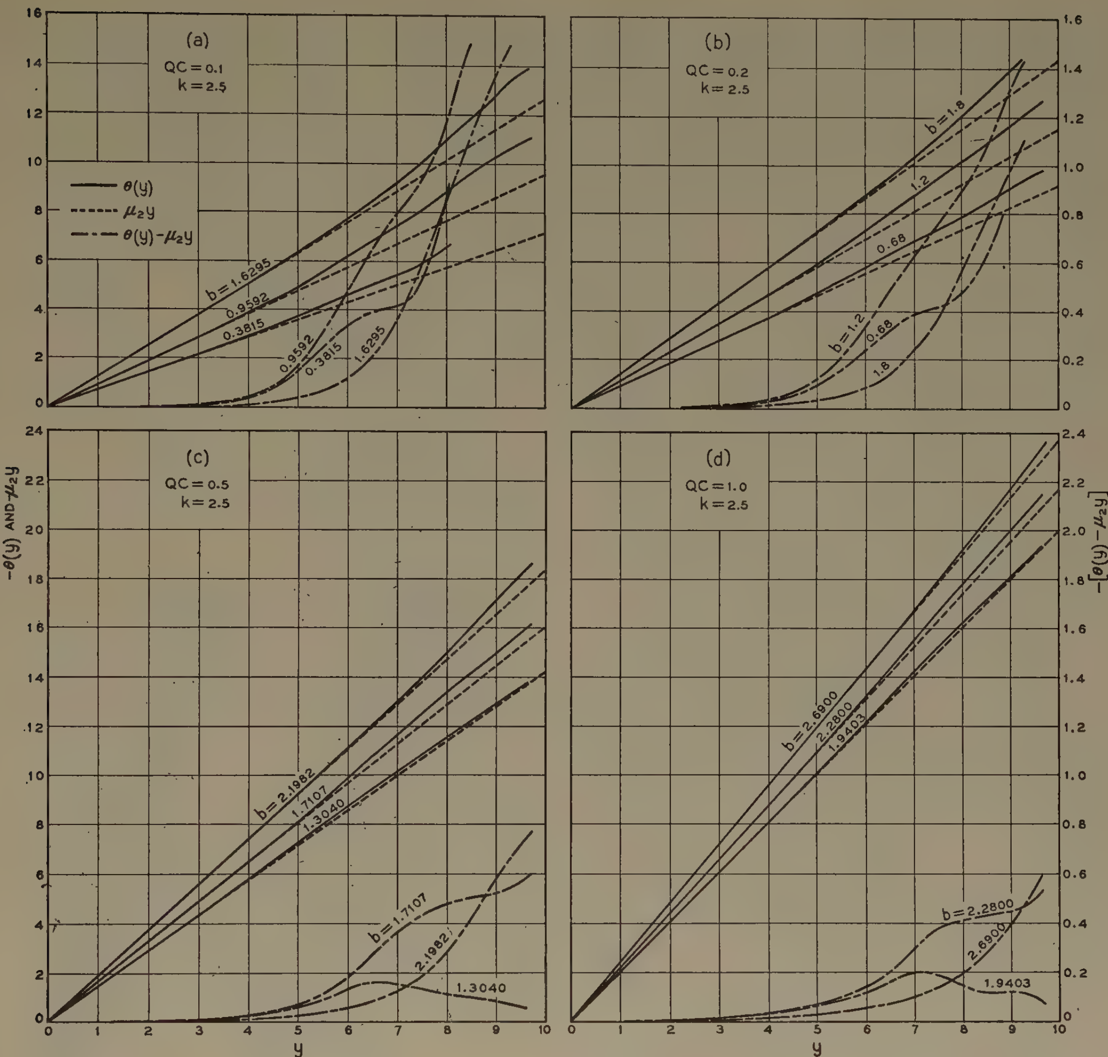


Fig. 6—$\theta(y)$, $\mu_2 y$, and $\theta(y)-\mu_2 y$ versus y curves for (a) cases (2), (3), and (4), (b) cases (7), (8), and (9), (c) cases (12), (13), and (14), (d) cases (17), (18), and (19).

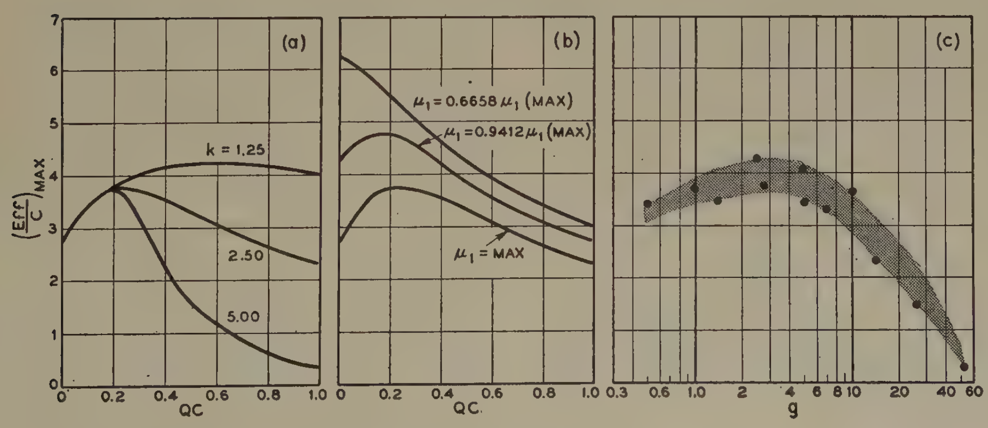


Fig. 7—(a) $(\text{Eff.}/C)_{\text{max.}}$ versus QC curves using k as the parameter, for the cases with b for maximum μ_1. (b) $(\text{Eff.}/C)_{\text{max.}}$ versus QC curves using the fraction of $\mu_{1(\text{max.})}$ as the parameter for the cases with $k=2.5$. The curves show how $(\text{Eff.}/C)_{\text{max.}}$ varies with b. (c) $(\text{Eff.}/C)_{\text{max.}}$ versus g curves for the cases with b for maximum μ_1.

This also provides a simple method to determine $i_1(z, t)$ experimentally, as we can easily measure $A(y)$ and $\theta(y)$ along the tube by means of a measuring probe. $|i_1(z,t)|/I_0$ is plotted in Fig. 8 for case (7). It has a maximum value of 1.24 at $y=8.0$. If the beam were perfectly bunched, $|i_1|/I_0$ would be equal to 2 as pointed out by Pierce.[2]

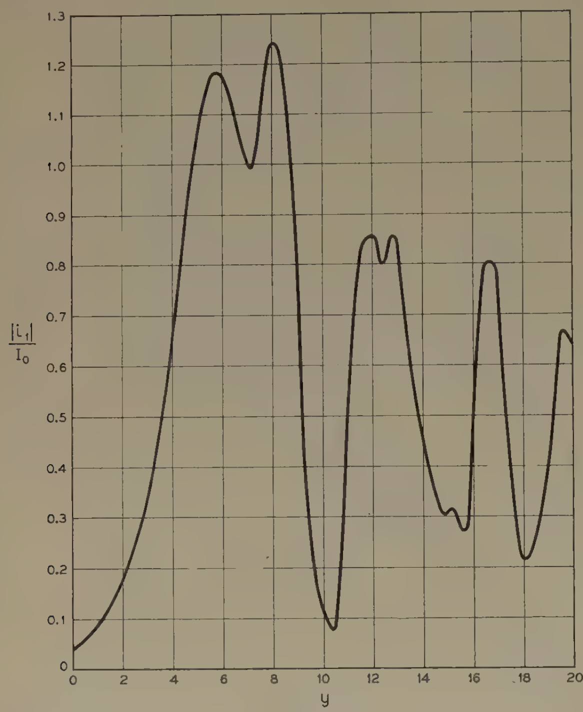


Fig. 8—($|i_1|/I_0$) versus y curve for case (7).

Before reaching the saturation level, $dA(y)/dy$ is always positive. $i_1(z, t)$ either leads or lags the circuit electric field by a phase somewhat between 90 and 180 degrees, depending upon the sign of the quantity

$$\left(\frac{d\theta(y)}{dy} + b\right).$$

The center of the electron bunching will be located somewhere between $\Phi=(2n+1.5)\pi$ and $(2n+2)\pi$, if

$$\left(\frac{d\theta(y)}{dy} + b\right)$$

is positive; somewhere $\Phi=(2n+1)\pi$ and $(2n+1.5)\pi$, if

$$\left(\frac{d\theta(y)}{dy} + b\right)$$

is negative, where n is an integer including zero. We shall return to this point later in the discussion related to Figs. 9(b), 9(c), and 9(d).

y Vs ($\Phi+\theta$) Curves—Applegate Diagrams

After having obtained a general picture of the circuit wave as discussed in a previous section, we shall next study the motions of the electrons and compare them for cases of different values of QC, k, and b. In the first place, we may gain insight into the beam behavior by studying $\Phi(\Phi_0, y)$'s and $q(\Phi_0, y)$'s at some interesting points along the tube, such as, the point at the saturation, the point at which the electrons start to overtake one another, etc. Further, we can analyze the interaction between the beam and the circuit by comparing $\Phi(\Phi_0, y)$'s and $q(\Phi_0, y)$'s at different values of y. Eventually we want to establish a dynamic picture of the electron motion by following the paths of the electrons and observing their velocities, and their phases with respect to the circuit field, as y increases. For these purposes, it is desirable to plot data in a form similar to the "Applegate Diagram," a sort of "space-time" diagram used to study klystrons. Figs. 9(a) to 9(h), below and on the following pages, are for cases (2), (7), (8), (9), (11), (12), (15), and (17), respectively.

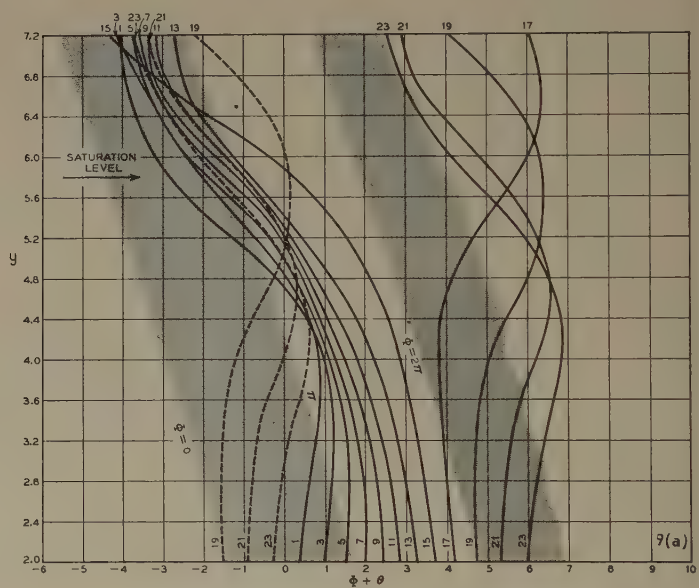


Fig. 9(a)—y versus ($\Phi+\theta$) curves for case (2).

In these figures, y versus ($\Phi+\theta$) curves are plotted with one curve for each electron. There should be 24 curves in an interval of 2π (in Φ or in $\Phi+\theta$), corresponding to the 24 electrons used in the calculations. Only odd numbered electrons and only electrons within an electronic wavelength (except few shown by dotted curves), however, are shown in the figures to avoid confusion by too many curves. It is understood that those curves should be repeated at every period of 2π (in Φ or in $\Phi+\theta$).

In the shaded regions (Fig. 9), the axially-directed circuit electric field is negative and accelerates the electrons in the positive z direction. Electrons are decelerated by the positive circuit field in the unshaded region. We will denote, for convenience, the shaded regions as the "accelerating regions" and the unshaded regions the "decelerating regions." The boundaries of these regions are constant Φ contours, from which we may determine the phase of the circuit wave at any position in this y versus $\Phi+\theta$ diagram.

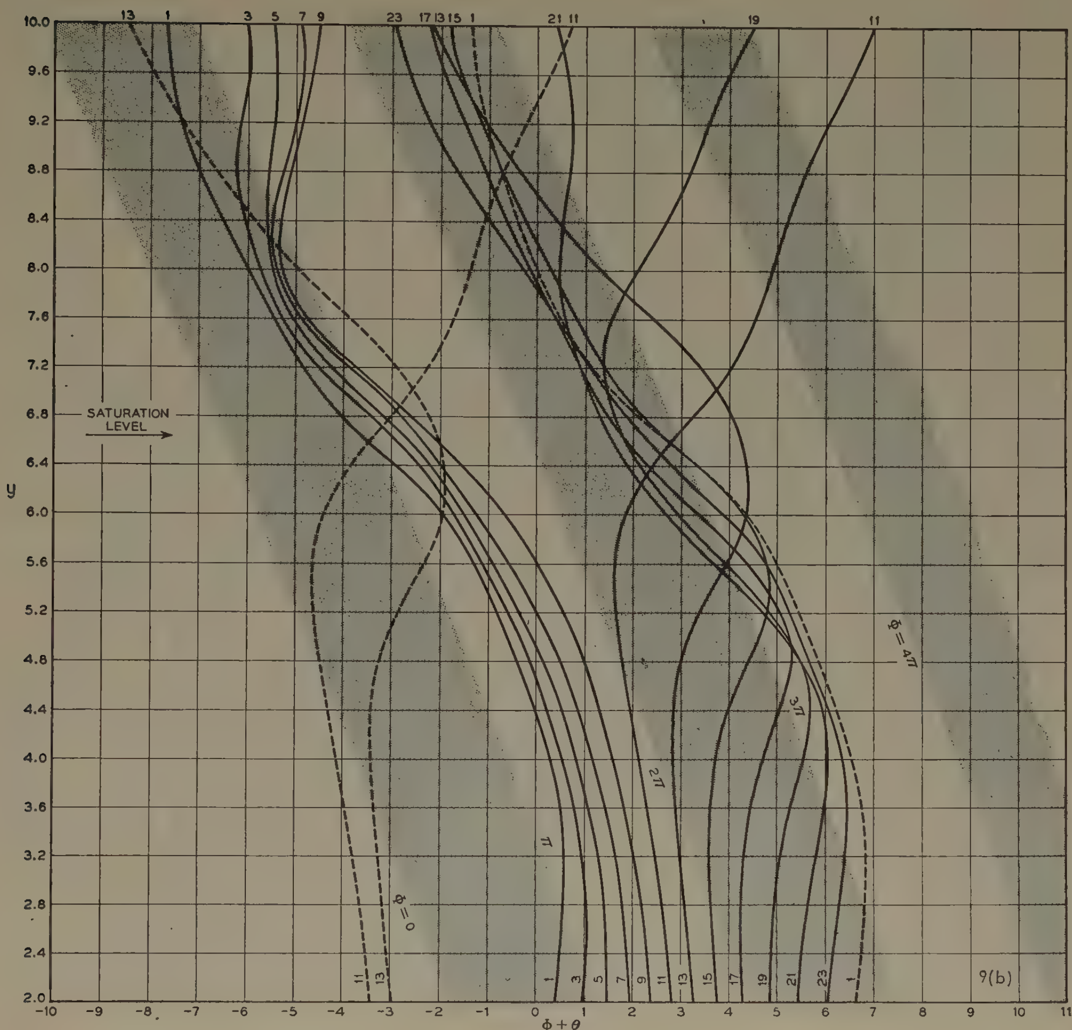


Fig. 9(b)—y versus $(\Phi+\theta)$ curves for case (7).

From (10) we have

$$\frac{\partial(\Phi+\theta)}{\partial y} = 2q(\Phi_0, y),$$

and thus the reciprocal of the slopes of the curves in Fig. 9 are proportional to the rf velocities of the electrons. As we have mentioned previously, a distribution of the electrons in Φ may be considered as the distribution in ωt or in y within a small interval of y. The distribution of the curves at any constant y line therefore represents the distribution of the electrons. More specifically, spacing between the curves should be inversely proportional to electron charge density, or electron current density when C is small. Concentration of curves indicates bunching of electrons in the beam.

Although the electron motions are drastically different for different cases, it is important to extract certain common features which we may consider as basic to the electron behavior. Let us take case (9) ($QC=0.2$, $k=2.50$, $b=1.8$) [Fig. 9(d)] as an example. At low level, the rf velocities are so small that the electron paths are almost vertical. However, electrons continuously advance in phase with respect to the circuit field, because their average speed is greater than the speed of the circuit wave. They will continue to do so up to a point at which some of the electrons are slowed down so much that they are trapped in the negative potential troughs of the field. In our case, this happens at y near 6. Up to this point, they behave much as in the small signal theory.

Suppose we are now in a co-ordinate system moving with the average electron speed u_0, in which $2q$ is electron velocity and $d\theta/dy$ is wave speed. It is obvious that electrons must attain a negative velocity comparable to the wave speed in order to stay with the circuit wave. For convenience, we shall divide the electrons into three groups, according to the types of the orbits that they have. At y near 6, electrons 7 to 19 have negative velocities because they are decelerated by the retarding circuit field in the region between $\Phi=3\pi$ and 4π. Among them, electrons 13 to 19 may be considered as of the first group. These have not built up enough negative velocity to follow the retrogression of the circuit wave and they will continue to advance ahead of the wave until they are trapped in the next decelerating region between $\Phi=5\pi$ and 6π. In the second group, however, electrons 7 to 11, which have velocities comparable to

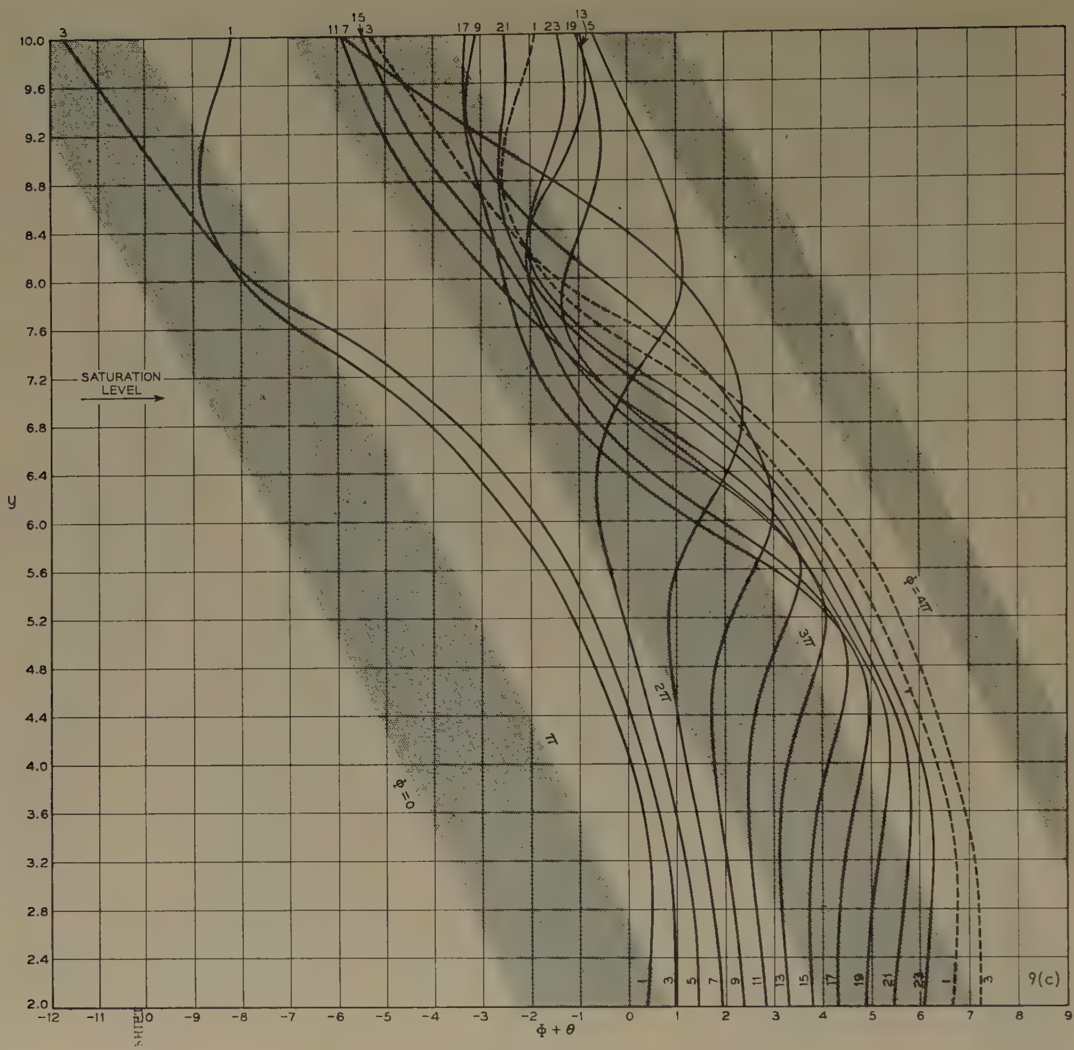


Fig. 9(c)—y versus $(\Phi+\theta)$ curves for case (8).

the wave speed, will follow paths almost parallel to the constant Φ contours. These electrons, which form the principal part of the bunch, will continuously travel inside the decelerating region and contribute power to the circuit until they cross the boundary, the $\Phi=3\pi$ line. After that, they enter the accelerating region, where they acquire the forward velocities. For the third group, we have electrons 1 to 5 and electrons 21 to 23 (shown as the dotted curves) of the later cycle. They have forward velocities at y near 6. Sooner or later, they enter the decelerating region where they are decelerated by both the retarding circuit field and the space-charge cloud of the electron bunch. They are finally pushed back to the accelerating region where they acquire forward velocities again. Some of them might miss the bunch and would then go straight forward. This, however, does not happen in this case. By extending y far enough, one will find that most of the electrons will be trapped in the negative potential troughs of the field and oscillate back and forth around the equilibrium positions, the $\Phi=(2n+1)\pi$ lines (n may be any integer). When they are in the decelerating region, they contribute power to the circuit, and in the accelerating region, they gain kinetic energy at the expense of the circuit power. We thus observe a series of the peaks in the $A(y)$ curves.

Although in general the electrons will follow the same types of motion as described above, for all the cases, it is interesting to study the effects of QC, k and b on their motions. To investigate these parameters separately, we plot y versus $\Phi+\theta$ diagrams for the following cases. Figs. 9(b), 9(c), and 9(d) are plotted for the cases having the same values of QC and k, but different electron injection speeds, that is, $QC=0.2$, $k=2.5$, and $b=0.68$, 1.2 and 1.8, respectively. Similarly, the cases of $QC=0.5$, $b=1.3040$, and $k=1.25$, 2.50 and 5.00 are shown in Figs. 9(e), 9(f), and 9(g), and the cases of $k=2.5$, b for maximum small-signal gain and $QC=0.1$, 0.2, 0.5, and 1.0 are in Figs. 9(a), 9(b), 9(f), and 9(h).

Considering the first set of figures; it is observed that when b is larger the wave speed $|d\theta(y)/dy|$ also increases. Electrons thus need greater rf velocities before they can follow the types of the orbits described above. As a matter of fact, the electrons will travel a longer distance before they reach the saturation level and the limiting power is larger even though μ_1 may be smaller.

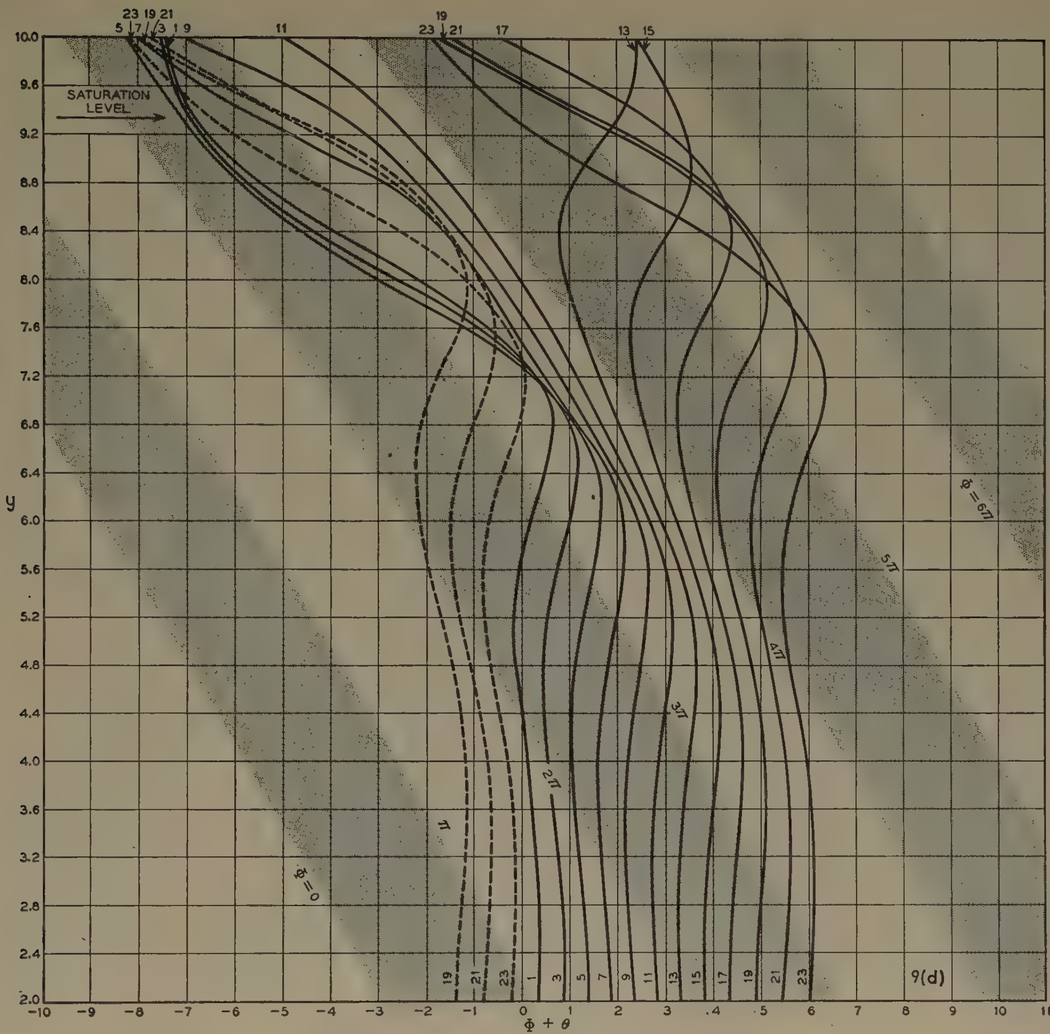


Fig. 9(d)—y versus $(\Phi+\theta)$ curves for case (9).

The effect of k can easily be seen by comparing Figs. 9(f) and 9(g). At low levels the electron orbits in both figures are identical. This should be so, as pointed out previously. At high levels, the electron bunching in Fig. 9(g) becomes loosened shortly after the electrons have started to overtake one another because of large short range space-charge forces, whereas in Fig. 9(b), the electron bunching stays tight even beyond the saturation level. The strong short-range space-charge forces generally distort the electron paths, particularly when the electrons approach one another in the bunching. This will reduce the limiting power or efficiency and the tube will saturate at a lower y. The effect of QC is interesting. When QC is large, the electron paths are spaced loosely and the rf velocities build up slowly. This, together with the fact that b is larger in the case in which QC is large, tends to raise the saturation level and to increase the limiting efficiency, as we have described, in connection with the effect of b. On the other hand, a large space-charge force, as in the case of large k, tends to distort the electron paths and to inhibit bunching, with the result of lowering the saturation level and limiting the efficiency. This explains the fact that efficiency generally increases with QC when QC or k is small, but is reduced sharply at larger values of QC and k.

In general, when b is small as in Fig. 9(b), the electron bunching starts in the left part of the decelerating region. A larger value of b tends to place the bunching toward the right. This can be seen in Figs. 9(c) and 9(d). The location of the bunches is related to $i_1(z, t)$, as has been mentioned. As y increases, the bunching moves gradually toward the left. At the saturation level, half of the bunch is located in the accelerating region and half in the decelerating region. The net power transfer between the circuit and the beam is then zero. In general, the beam contains a strong fundamental component at the saturation level. The bunching is not effective simply because of the position of the bunch with respect to the circuit field. Thus we may increase the efficiency by introducing a sudden phase-shift to the circuit wave near the saturation level such that the bunch may be moved well inside the decelerating region. The electrons in the bunch will then continue to contribute power until they reach the accelerating region again.

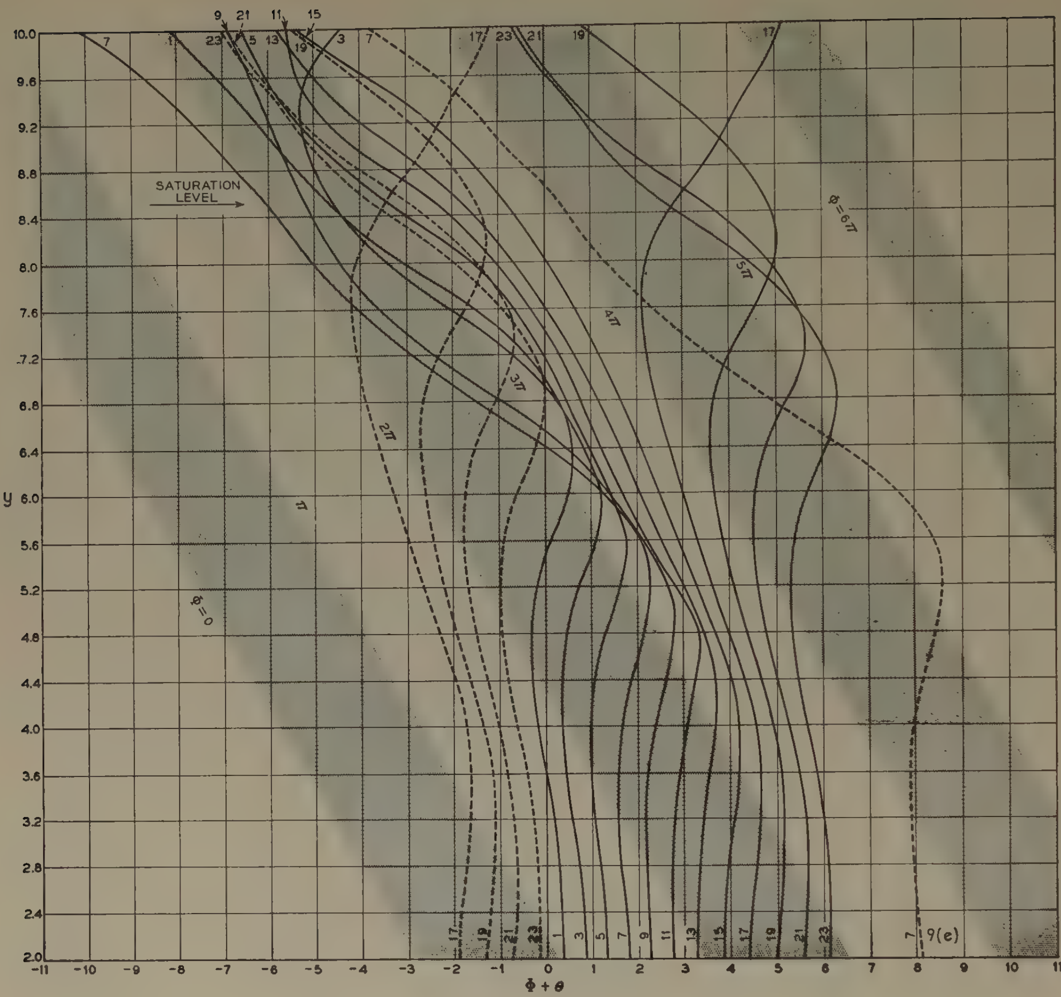


Fig. 9(e)—y versus $(\Phi+\theta)$ curves for case (11).

Acknowledgment

The authors wish to thank Dr. J. R. Pierce and C. C. Cutler for many valuable suggestions during the course of this investigation. The numerical computations were carried out by Miss D. C. Leagus. The accuracy, ingenuity and speed consistently applied by her to the solution of this problem have been invaluable, and the authors wish to express their sincere appreciation of her excellent contribution. The experience of P. R. Peabody, who programmed the problem for an IBM Card Programmed Calculator in 1953, was most helpful in the present calculations.

Appendix I

Put

$$V(z, t) = \tfrac{1}{2} w_1 e^{j[\omega t-(\omega/u_0)z]} + \tfrac{1}{2} w_1^* e^{-j[\omega t-(\omega/u_0)z]}$$

where

$$w_1 = A(z)e^{j\theta(z)} \tag{29}$$

and

$$X(\Phi_0, y) = \int_{y(\Phi_0,0)}^{y} 2q(\Phi_0, y)dy.$$

By means of (10) this is obviously

$$X(\Phi_0, y) = \Phi(\Phi_0, y) - \Phi_0 + \theta(y). \tag{30}$$

Eqs. (7), (8), (9), and (10) can be reduced to the form

$$\frac{dw_1}{dy} + jbw_1 = \frac{j}{2\pi}\int_0^{2\pi} e^{j[X+\Phi_0]}d\Phi_0 \tag{31}$$

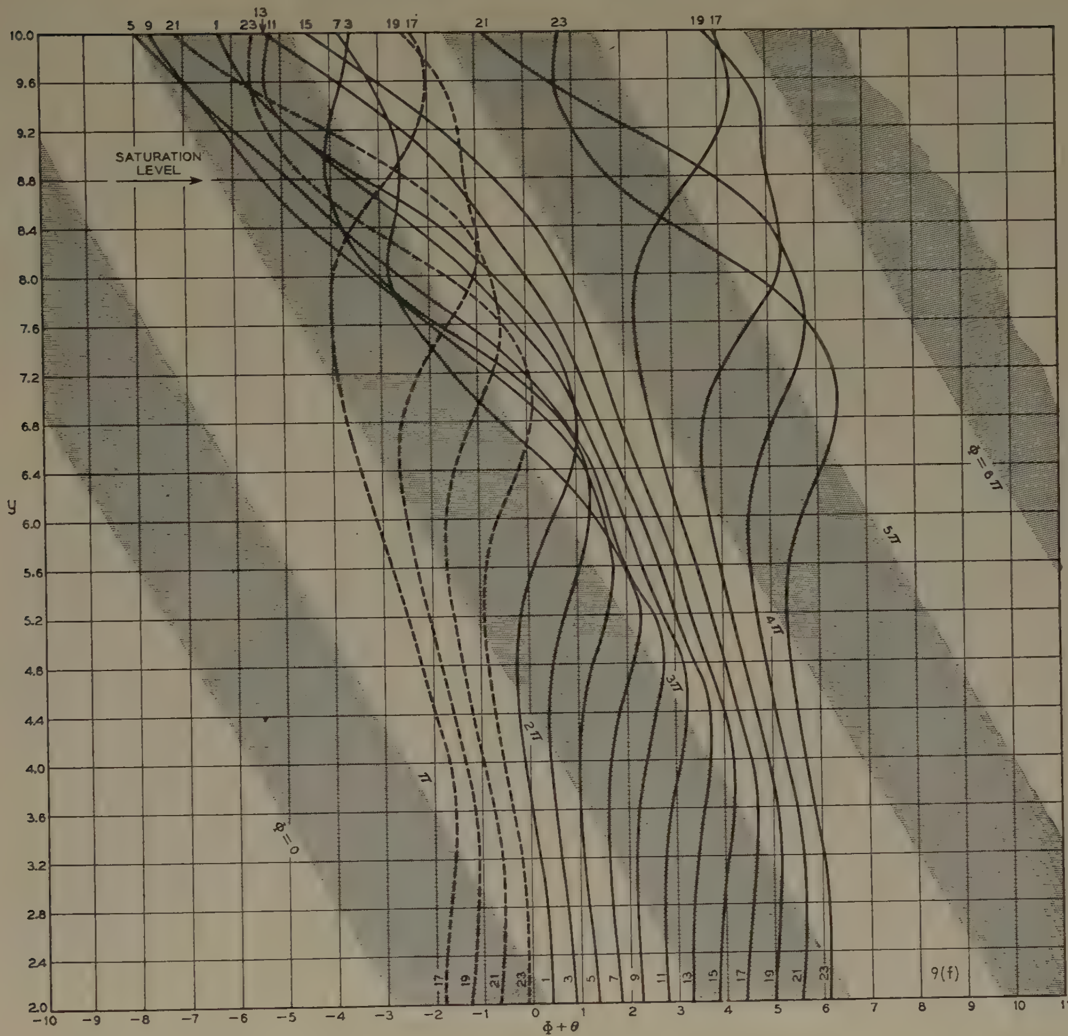
$$\frac{\partial^2 X(\Phi_0, y)}{\partial y^2} = jw_1 e^{-j[X+\Phi_0]} - jw_1^* e^{j[X+\Phi_0]} - \frac{e}{m\omega^2 C^2}\frac{I_0}{u_0}\int_{-\infty}^{+\infty} B\left[\frac{u_0}{\omega}\{\Phi(\Phi_0+\phi, y) - \Phi(\Phi_0, y)\}\right]d\phi. \tag{32}$$

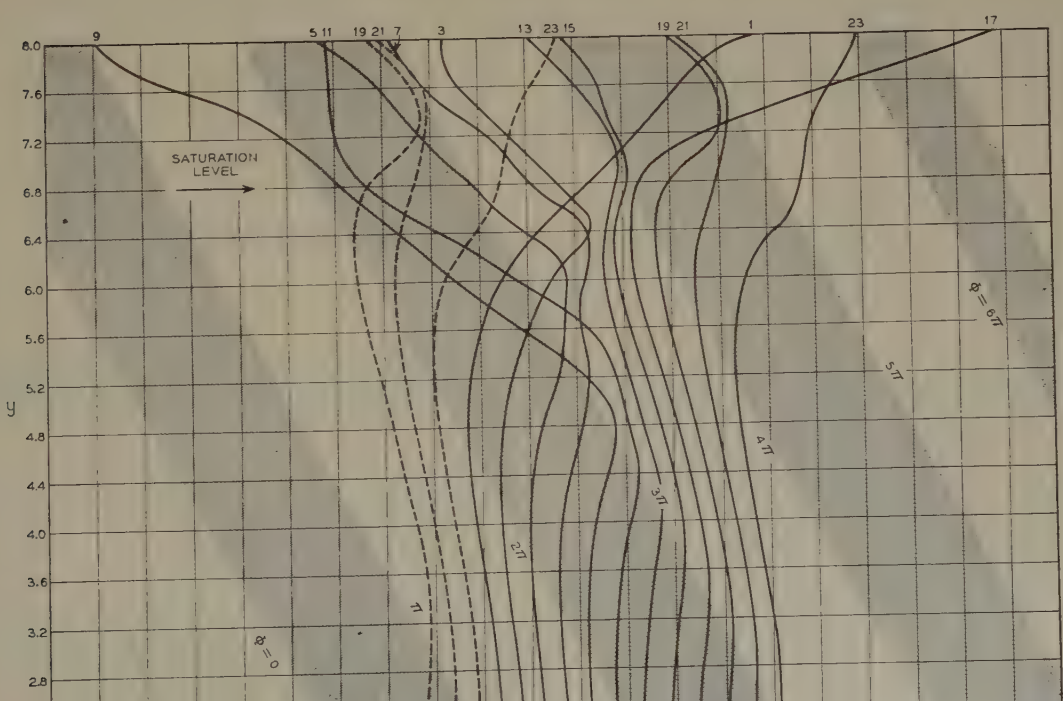
In the linear theory, we may put

$$X = pe^{-j\Phi_0} + p^* e^{j\Phi_0}.$$

The space charge integral in (32) then becomes

$$\int_{-\infty}^{+\infty} B'\left(\frac{u_0}{\omega}\phi\right)\Delta X d\phi = \int_{-\infty}^{+\infty} B'\left(\frac{u_0}{\omega}\phi\right) \cdot [pe^{-j\Phi_0}(e^{-j\phi}-1) + p^* e^{j\Phi_0}(e^{j\phi}-1)]d\phi, \tag{33}$$





Figs. 9(f) and (g)—y versus $(\Phi+\theta)$ curves for cases (12) and (15), respectively.

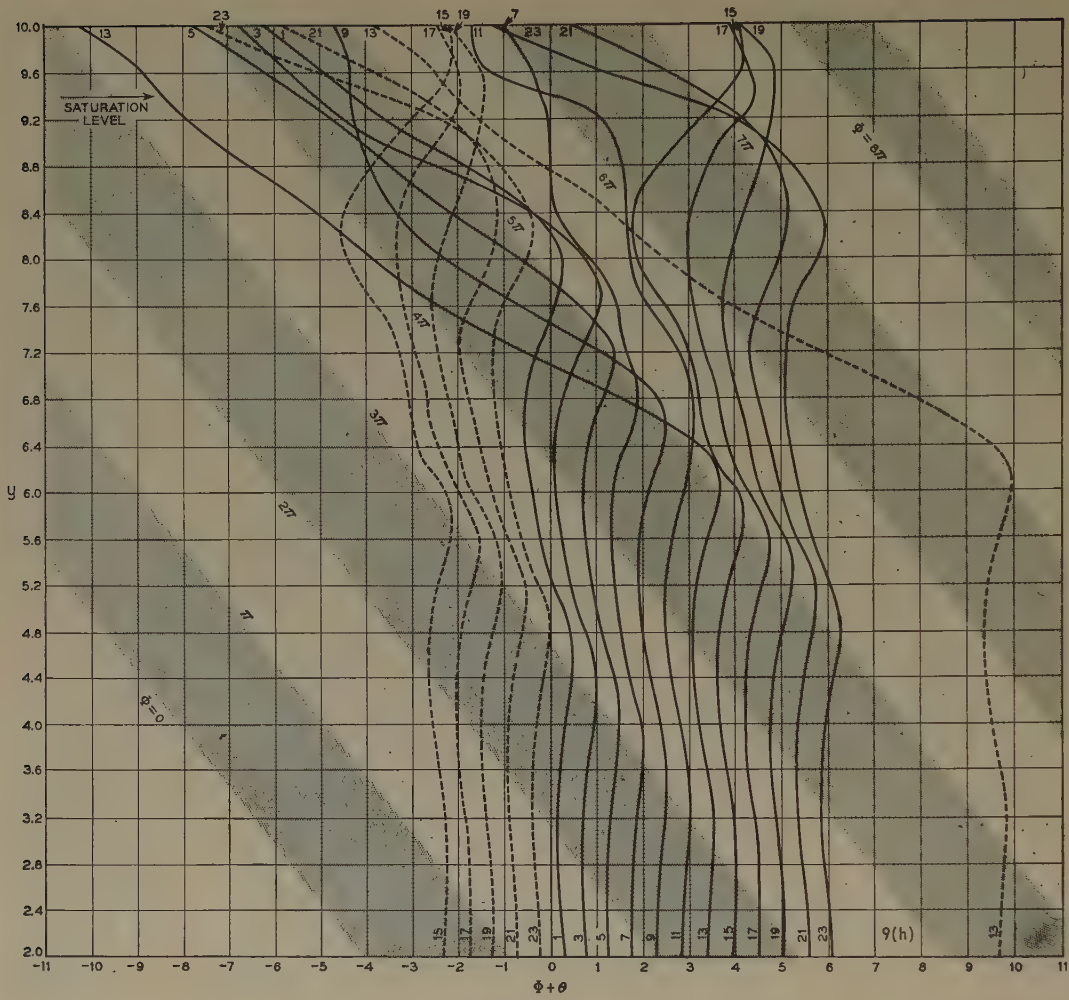


Fig. 9(h)—y versus $(\Phi+\theta)$ curves for case (17).

where $B'[(u_0/\omega)\phi]$ is the derivative of $B(z)$ with respect to ϕ. Expanding

$$e^{iX} = 1 + jX + \cdots$$

we have for (31) and (32), respectively,

$$\frac{dw_1}{dy} + jbw_1 = \frac{i}{2\pi}\int_0^{2\pi} e^{i\Phi_0}[1 + jX]d\Phi_0 = -p \tag{34}$$

$$\frac{d^2p}{dy^2} = jw_1 - p\frac{e}{m\omega^2C^2}\frac{I_0}{u_0}\int B'\left(\frac{u_0}{\omega}\phi\right)(e^{-i\phi} - 1)d\phi. \tag{35}$$

Substituting $-p$ in (34) into (35), we obtain

$$\left(\left[\frac{d^2}{dy^2} + \frac{e}{m\omega^2C^2}\frac{I_0}{u_0}\int B'\left(\frac{u_0}{\omega}\phi\right)(e^{-i\phi} - 1)d\phi\right]\cdot\left[\frac{d}{dy} + jb\right] + j\right)w_1 = 0. \tag{36}$$

Comparing this with the Pierce equation,[8]

$$(\delta^2 + 4QC)(\delta + jb) + j = 0 \tag{37}$$

[8] Pierce, *op. cit.*, eq. (7.13).

we find,

$$4QC = \frac{e}{m\omega^2C^2}\frac{I_0}{u_0}\int B'\left(\frac{u_0}{\omega}\phi\right)(e^{-i\phi} - 1)d\phi. \tag{38}$$

According to (19), we have finally (22) or (23) of the text.

Appendix II

The following is the second-order small-signal theory used in this paper to calculate the initial conditions. Put

$$V(z, t) = \tfrac{1}{2}ve^{i[\omega t-(\omega/u_0)z]} + \tfrac{1}{2}v^*e^{-i[\omega t-(\omega/u_0z)]}$$

$$v = \epsilon v_1 + \epsilon^2 v_2 \tag{39}$$

$$\chi = \epsilon\chi_1 + \epsilon^2\chi_2. \tag{40}$$

Here ϵ is a small quantity and χ is X defined in (30) of Appendix I. The following quantities may be expanded into power series in ϵ.

$$e^{i\chi} = 1 + \epsilon j\chi_1 + \epsilon^2\left(j\chi_2 - \frac{\chi_1^2}{2}\right) + \cdots \tag{41}$$

$$e^{-j\chi} = 1 - \epsilon j\chi_1 + \epsilon^2\left(-j\chi_2 - \frac{\chi_1^2}{2}\right) + \cdots \quad (42)$$

$$B\left[\frac{u_0}{\omega}\{\Phi(\Phi_0+\phi, y) - \Phi(\Phi_0, y)\}\right] = B\left[\frac{u_0}{\omega}(\phi + \Delta\chi)\right]$$
$$= \left[B'\left(\frac{u_0}{\omega}\phi\right)\Delta\chi\right]$$
$$+ \epsilon^2\left[B'\left(\frac{u_0}{\omega}\phi\right)\Delta\chi_2 + \frac{B''\left(\frac{u_0}{\omega}\phi\right)}{2}(\Delta\chi_1^2)\right]$$
$$+ \cdots, \quad (43)$$

where B' and B'' are respectively the first and the second derivatives of B with respect to ϕ. By putting

$$\chi_1 = pe^{-j\Phi_0} + p^*e^{j\Phi_0} \quad (44)$$
$$\chi_2 = qe^{-2j\Phi_0} + \gamma + q^*e^{2j\Phi_0}. \quad (45)$$

The equations of the first-order small quantities can be derived from (7) to (10) and are given by (34) and (35) in Appendix I, with $w_1 = v_1$. The equations of the second-order small quantities are

$$\frac{dv_2}{dy} + jbv_2 = \frac{j}{2\pi}\int_0^{2\pi} e^{j\Phi_0}\left(j\chi_2 - \frac{\chi_1^2}{2}\right)d\Phi_0 \quad (46)$$

$$\frac{d^2\chi_2}{dy^2} = je^{-j\Phi_0}(-j\chi_1 v_1) - je^{j\Phi_0}(j\chi_1 v_1^*)$$
$$- \frac{e}{m\omega^2C^2}\frac{I_0}{u_0}\int_{-\infty}^{+\infty}\left[B'\left(\frac{u_0}{\omega}\phi\right)\Delta\chi_2\right.$$
$$\left.+ \frac{B''\left(\frac{u_0}{\omega}\phi\right)}{2}(\Delta\chi_1)^2\right]d\phi. \quad (47)$$

From (46), v_2 must be zero, as χ_2 and χ_1^2 do not contain any terms involving $e^{j\Phi_0}$, or its conjugate $e^{-j\Phi_0}$. We may thus put

$$v = \epsilon v_1 = \epsilon e^{\mu y} \quad (48)$$
$$\mu = \mu_1 + j\mu_2, \quad (49)$$

where μ_1 and μ_2 are Pierce's[2] x_1 and y_1 respectively. By means of (44) to (49), we obtain the following relations after considerable algebra,

$$\chi = -2\epsilon(|\mu+jb|\, e^{\mu_1 y}\cos[\arg(\mu+jb)+\mu_2 y-\Phi_0])$$
$$+\epsilon^2\left(-2|N|\, e^{2\mu_1 y}\cos[\arg N+2\mu_2 y-2\Phi_0]-\frac{e^{2\mu_1 y}}{2\mu_1}\right) \quad (50)$$

$$2q = -2\epsilon(|\mu(\mu+jb)|\, e^{\mu_1 y}\cos[\arg\mu(\mu+jb)+\mu_2 y-\Phi_0])$$
$$+\epsilon^2(-4|\mu N|\, e^{2\mu_1 y}\cos[\arg\mu N+2\mu_2 y-2\Phi_0]-e^{2\mu_1 y}) \quad (51)$$

$$\frac{e}{u_0 m\omega C^2}E_s = -\epsilon 8QC|\mu+jb|\, e^{\mu_1 y}\cos[\arg(\mu+jb)+\mu_2 y-\Phi_0]$$
$$+\epsilon^2|M|\, e^{2\mu_1 y}\cos[\arg M+2\mu_2 y-2\Phi_0], \quad (52)$$

where

$$N = (\mu+jb)\frac{1+j12QC(\mu+jb)\dfrac{k^2}{k^2+4}}{4\mu^2+\dfrac{16QC(k^2+1)}{k^2+4}} \quad (53)$$

$$M = \frac{8QC}{k^2+4}\left[\frac{(2k^2-1)-j12k^2QC(\mu+jb)}{\mu^2+4QC\dfrac{k^2+1}{k^2+4}}\right](\mu+jb). \quad (54)$$

The initial conditions at $y=0$ are computed from (50) to (52) by setting $\epsilon = 0.03$ for all the cases. For this value of ϵ, the maximum value of $|2q|$ is less than 10 per cent of $|d\theta/dy|$ for the worst case. Φ is calculated from χ by (30).

Saturation Current in Alloy Junctions*

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Summary—According to theory, the current which flows across a p-n junction when it is biased in the reverse direction should be nearly independent of the applied voltage. This current is called the saturation current, I_s. In actual practice, the current often increases with voltage (due to leakage paths across the junction at the surface) and eventually large currents flow as the breakdown voltage of the junction is approached. However, for "good" junctions well below their breakdown voltage, the reverse current follows theory so that the magnitude of I_s is of considerable practical interest.

A previous equation[1] for saturation current has been given which applies where the effect of free surfaces may be ignored and where there are quite thick (greater than a diffusion length) layers of material on either side of the junction. For alloy junctions made on thin pieces of semiconductor, these conditions are violated.

In this paper a new equation is developed which applies specifically to diodes made by alloying circular junctions on thin wafers. Over its range of applicability, it gives accurate values of I_s. It is shown that most of the saturation current comes from thermal generation at the free surfaces of the base wafer. The equation for I_s consists of a geometrical term and a coefficient which depends on the physical constants of the material. The principal dependences are these: I_s increases linearly with base wafer resistivity and exponentially with temperature. It also increases, but more slowly, with wafer thickness and surface recombination velocity.

For the collector of an alloy junction transistor, the basic equation is the same as for a diode, except for a small correction involving the emitter area. For the emitter junction, reverse saturation current is often not of great interest but can also be calculated; this requires a modification of the basic equation which is also given in this paper.

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[1] W. Shockley, "Holes and Electrons in Semiconductors," D. Van Nostrand Co., Inc., New York, N. Y., p. 314; 1950.

General Discussion

Saturation current is due to thermal generation of hole-electron pairs on both sides of a p-n junction. The pairs diffuse toward the junction where they are separated by the applied field so that current flows. Generation occurs at surfaces and in the volume just as does recombination. This is not surprising since the two phenomena are intimately connected. Under equilibrium conditions, holes and electrons are constantly generated and recombining. These two processes must take place, on the average, at equal rates. Otherwise, the hole and electron concentrations would eventually become either zero or infinite. From measurements of the lifetime of excess hole-electron pairs (produced, for example, by illuminating a sample), the rate of recombination at equilibrium can be computed. This is also the rate of generation at equilibrium. The net generation is defined as the difference between this total generation and whatever recombination may be taking place.

Analytically, generation may be looked upon as "negative" recombination and treated with the same mathematical techniques. For example, in n-type material, the volume recombination rate per unit volume is commonly given as $(p-p_0)/\tau_v$ where p is the hole density (minority carriers), p_0 is the equilibrium value of p, and τ_v is the average hole lifetime for holes lost through volume recombination. When p is less than p_0, the recombination rate becomes negative and, in fact, represents the net rate of thermal generation. Similarly, surface recombination and generation are both represented by $(p-p_0)s$ (per square cm second), where s is the surface recombination velocity. The saturation current, I_s, is calculated by integrating these terms over the volume and surface which contribute reverse current. In the present work, calculations are made for an n-type base wafer, such as would be used for a p-n-p alloy junction transistor. The same form applies to p-type material by appropriate exchange of symbols.

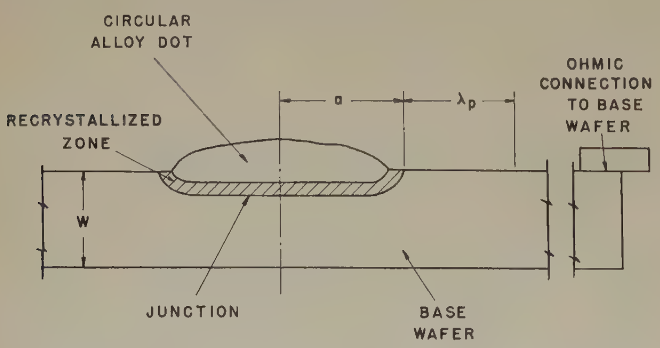

Fig. 1—Alloy junction diode geometry consisting of circular alloy dot on thin base wafer.

Saturation Current in Diodes

Consider the diode geometry in Fig. 1 above. Four regions contribute to I_s: (1) the volume of n-type wafer within a hole diffusion length, λ_p, of the junction; (2) the wafer surfaces within the same distance, λ_p; (3) the volume of the recrystallized material within an electron diffusion length, L_e, in the p-type material; and (4) the surface of the p-type material within L_e.

λ_p is determined by the actual hole lifetime in the n-type wafer which, in the present case, is almost totally controlled by surface recombination. As such, it differs from the usual diffusion length (generally given the symbol L_p) which is determined only by the volume lifetime and applies where surface effects may be neglected. To avoid confusion and emphasize the difference, λ_p is used herein to designate the surface controlled diffusion length. If the actual hole lifetime is called τ_m, then λ_p is given by the expression

$$\lambda_p = \sqrt{D_p \tau_m}, \tag{1}$$

where D_p is the diffusion coefficient for holes in the n-type base material. For thin wafers such as those used for alloy junction devices,

$$1/\tau_m = 1/\tau_v + \frac{2s}{W}, \tag{2}$$

where τ_v is the mean lifetime of holes lost by volume recombination, s is the surface recombination velocity and W is the wafer thickness.[2] Usually, the most convenient way to compute λ_p is from direct measurement of τ_m made on the alloy junction sample itself. This may be done by measuring the transient response of the junction diode.[3,4]

For simplicity, the analysis is limited to the case where $2s/W \gg 1/\tau_v$. This means that volume recombination and generation are neglected. Typical values of s and τ_v are 300 cm/second and 200 μsec, respectively, in germanium. The wafer thickness, W, is usually less than 2.5×10^{-2} cm (0.010 inch). For such conditions, the contribution from the volume is less than 20 per cent of the contribution from the surface so that the approximation is not a serious limitation. Now, λ_p may be written as

$$\lambda_p = \sqrt{\frac{WD_p}{2s}}. \tag{3}$$

In addition, the two contributions to I_s from the recrystallized region are of negligible importance in practice as can be shown by the following argument: The ratio of electron current, I_n, which flows from the alloy region to the base wafer, to the hole current, I_p, which flows in the opposite direction, is given by the familiar relation

$$\frac{I_n}{I_p} = \frac{\sigma_b \lambda_p}{\sigma_e L_e}, \tag{4}$$

[2] Eq. (2) is derived by combining (17) and (19b) appearing on pp. 322 and 323 of Reference 1. It applies only where $sW \ll D_p$, which limits the analysis to the case of thin wafers.

[3] S. Lederhandler and L. J. Giacoletto, "Measurement of Minority Carrier Lifetime and Surface Effects in Junction Devices," to be published in Proc. I.R.E.

[4] B. R. Gossick, "Post-injection barrier electromotive force of p-n junction," *Phys. Rev.*, vol. 91, p. 1012; 1953.

where σ_b and σ_e are the conductivities of the base and p-type regions, respectively. Eq. 4 is related to the injection efficiency of the junction. From measured values of emitter efficiency of p-n-p alloy junction transistors, one may compute values of I_n/I_p which are generally much less than 0.1.

Thus, the only important contribution to I_s is from the surface of the base wafer. An approximate expression for I_s results from assuming that all holes generated at the free surfaces bounded by the radii a and $(a+\lambda_p)$ are collected and that the number of holes generated per second per square centimeter of free surface within this range is $p_0 s$. Then,

$$I_s = qp_0 s[2\pi(a+\lambda_p)^2 - \pi a^2], \tag{5}$$

where q is the electronic charge.

Eq. 5 may be put into more familiar terms by replacing s with $D_pW/2\lambda_p^2$ and p_0 with

$$p_0 = \frac{\rho}{\rho_i^2}\cdot\frac{b}{(1+b)^2}\cdot\frac{kT}{q^2D_p}.$$ [5]

Here, ρ is the resistivity of the base wafer, ρ_i the resistivity of the intrinsic semiconductor, and b is the ratio of electron to hole mobilities. The result is

$$I_s = \frac{kT}{q}\frac{\rho}{\rho_i^2}\frac{b}{(1+b)^2}\pi W\left[1+\frac{2a}{\lambda_p}+\frac{1}{2}\left(\frac{a}{\lambda_p}\right)^2\right], \tag{6}$$

where

$$\lambda_p = \sqrt{D_p\tau_m} \approx \sqrt{\frac{D_pW}{2s}}.$$

ρ_i decreases exponentially with temperature so that I_s increases rapidly. In germanium, the rate of increase of I_s is approximately 10 per cent per degree C. in the neighborhood of room temperature.

Eq. 6 may be evaluated for the case of an n-type germanium wafer where $b=2$, $\rho_i=47$ ohm-cm at 25 degrees C., and $D_p=44$ cm²/second:

$$I_s = 7.6We^{\Delta T/10}\left[1+\frac{2a}{\lambda_p}+\frac{1}{2}\left(\frac{a}{\lambda_p}\right)^2\right] \text{ microamperes,}$$

and

$$\lambda_p = 4.7\sqrt{\frac{W}{s}} \text{ cm.}$$

ΔT is the difference between operating temperature and 25 degrees C. The exponential in ΔT accounts for the dependence of ρ_i on temperature. For a typical case where $\rho=4$ ohm cm, $W=0.013$ cm, $a=0.06$ cm, and $s=300$ cm/second, $\lambda_p=0.03$ cm, and $I_s=2.8$ μa at 25 degrees C.

Eq. 6 may also be applied to p-type base material by substituting $\lambda_n=\sqrt{D_n\tau_m}$ for λ_p, where D_n is the electron diffusion coefficient. Since $b/(1+b)^2$ has the same value if $1/b$ is substituted for b (i.e., when hole and electron mobilities are interchanged), no other changes need be made. In germanium λ_n will be greater than λ_p by about 40 per cent for the same value of τ_m. Thus, for the same resistivity, I_s should be a little lower in p-type base material than in n-type.

The analysis was initially limited to the case where $W \ll 2s\tau_v$ so that volume generation is neglected. For thicker wafers, (or smaller s or τ_v), the contribution from the volume must be included. This is automatically accomplished if λ_p is computed from lifetime measurements made on the sample itself, using (1) as suggested above.

In typical germanium diodes of about five mils thickness, λ_p is in the neighborhood of ten mils. For "a" less than $4\lambda_p$ (0.040 inch) the term $(1+2a/\lambda_p)$ dominates the geometrical factor. For large junctions, the term $a^2/2\lambda_p^2$ dominates. Since λ_p^2 is approximately proportional to W, this means that I_s increases with wafer thickness more rapidly for small junctions than for large ones. Also, in small junctions I_s increases approximately as $\sqrt{s}$ while the dependence approaches the first power for large junctions.

A more rigorous calculation of the geometrical terms involves integration of the net generation, $(p_0-p)s$, over the surface. We may make the assumption that $p\approx 0$ on the surface region opposite the junction when W is small but in the circular ring where $a<r<(a+\lambda_p)$, strong radial hole density gradients exist and this assumption is certainly untrue. Rightfully, the terms $(1+2a/\lambda_p)$ should be replaced by the solution of the equations

$$J_p(r) = -qD_p\frac{dp}{dr} \tag{7}$$

and

$$2qs\int_\infty^r [p_0 - p(r)]rdr = -rWJ_p(r). \tag{8}$$

The first of these equations simply states that holes flow in the wafer by diffusion. The second demands continuity of hole current by equating the total hole current flowing toward the junction through a circle of radius r to the net generation from that point outward. These equations may be combined to give a Bessels' equation of zero order whose solution has an imaginary argument:

$$\frac{1}{r}\frac{d}{dr}\left(r\frac{dy}{dr}\right) = \frac{2s}{WD_p}y = \frac{y}{\lambda_p^2}, \tag{9}$$

where $y=p_0-p$.

Without going into further detail, the term needed to replace $(1+2a/\lambda_p)$ in (6) turns out to be

$$f\left(\frac{a}{\lambda_p}\right) = \frac{2a}{\lambda_p}\frac{-H_1^{(1)}(ia/\lambda_p)}{iH_0^{(1)}(ia/\lambda_p)}. \tag{10}$$

[5] The origin of this expression is clear from Reference 1. Shockley's expression for saturation current and the result of the present derivation have the same coefficient; they differ in the geometrical terms.

The terms $-H_1^{(1)}$ and $iH_0^{(1)}$ are tabulated functions[6] and $f(a/\lambda_p)$ is plotted, with $(1+2a/\lambda_p)$, against a/λ_p in Fig. 2. Only for a/λ_p less than about 0.3 do the two solutions depart appreciably. In most practical cases a/λ_p is greater than 0.3 so that the approximate expression is sufficiently good. If $a<W$, (6) will give values of I_s which are somewhat high since density gradients across the wafer thickness (which were neglected in the calculation) will become appreciable. Within the usual range of application, however, this is not a likely source of error.

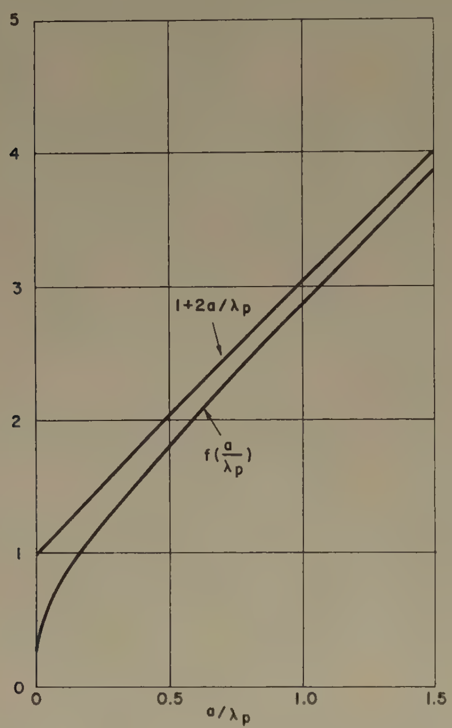

Fig. 2—Approximate solution $(1+2a/\lambda_p)$ compared to Bessel function solution $f(a/\lambda_p)$.

Experimental Verification

In order to verify (6), I_s was measured and computed for a series of 48 alloy junction diodes of different geometry made on n-type germanium wafers of various resistivities. λ_p was computed from τ_m measurements according to (1). The bulk resistivity was measured on the original ingots and also checked for local variations in the following manner: The capacitance of the junction when biased in the reverse direction by a voltage, V, is

$$C \approx 4 \times 10^{-8} \frac{a^2}{\sqrt{V\rho}} \text{ farads.}$$

"a" and C were determined by direct measurement (traveling microscope and capacitance bridge respectively). From these, ρ was computed. In this experiment, I_s varied from 0.8 to 23 μamps, ρ from 0.5 to 15 ohm cm, τ_m from 1 to 30 microseconds, and "a" from 10 to 53 mils. All samples were five mils in thickness. About 80 per cent of the measured values fell within ± 50 per cent of the calculated values and the average of the ratio of calculated to measured values is within one per cent of unity. This agreement is even better than one would expect considering the possible errors. The accuracy of measurement of resistivity alone would account for the observed spread.

Saturation Current in Transistors

The above development applies principally to diodes since it assumes a free surface opposite the dot. For the collector of a transistor whose emitter is floating, the emitter area should be subtracted out of the geometrical term. This amounts to replacing $(a/\lambda_p)^2$ with $(a^2-c^2)/\lambda_p^2$ in (6), where c is the radius of the emitter. This is usually a small correction since the emitter area is much less than the collector area. For the emitter, however, (6) gives much lower values of I_s than those one measures. This is because the floating collector is able to enlarge the collecting ability of the emitter by what is called the feed-in feed-out effect. By this mechanism, holes are collected by the collector near its periphery and translated to its center where they are re-emitted to contribute to the emitter saturation current. Thus, the emitter saturation current is related to the collector saturation current. An approximate expression for the emitter saturation current, I_{es}, in terms of the collector saturation current, I_{cs}, (where both currents are measured on same transistor with opposite junction floating and collector substantially larger than emitter), is:

$$I_{es} = \frac{I_{cs}}{1 + \frac{WW_b}{c^2}\left[1 + \frac{2a}{\lambda_p} + \frac{1}{2}\left(\frac{a}{\lambda_p}\right)^2\right]}, \qquad (11)$$

where W_b is the average emitter-collector spacing, "c" is the emitter radius and "a" the collector radius as before. For example, when $a=0.060$ cm, $c=0.020$ cm, $W=0.013$ cm, $W_b=0.005$ cm, and $\lambda_p=0.030$ cm, I_{es}/I_{cs} should equal 0.67. For a transistor of this geometry, the experimental value was 0.69, suggesting that (11) is valid for usual ratio of emitter to collector areas.

[6] E. Jahnke and F. Emde, "Tables of Functions," G. E. Stechert and Co., New York, N. Y.; 1938.

Theory of Radio Transmission by Tropospheric Scattering Using Very Narrow Beams*

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Summary—Calculations have been made for communication over a 300-km path between antennas, each consisting of a paraboloid of diameter 100 λ. It is assumed that, under normal atmospheric conditions, transmission over this distance is due to scattering by atmospheric irregularities and that the scattering phenomena are described by the formula used by Booker and Gordon.[1] Over this path, the effect of spread in the direction of arrival of scattered power should become noticeable for beamwidths less than about 1.5 degrees. For the 0.73 degree-beams assumed, spread in the angle of arrival should show up quite markedly for synchronized beam swinging—that is, when the beams at both transmitter and receiver are swung simultaneously so that their axes continue to intersect. When the beams are swung 1 degree to one side of the great circle path, the reduction in power received should be about 7 db, as compared with more than 40 db that would occur for propagation purely in the vertical plane containing transmitter and receiver. Narrowing the antenna beams to very small values is an important key to increasing circuit radio-frequency bandwidth capability, although at the expense of increasing antenna "aperture-medium coupling loss" (so-called "gain loss"). For the assumed circuit average bandwidth should be about 6 mcp, the loss being of the order of 10 db. For beams much larger than 1.5 degrees, the bandwidth should be limited by the medium to about 3 mcp, with negligible loss.

General formulas are given from which the important characteristics of any such communication link can be predicted.

I. Introduction

It is the object of this paper to make some calculations about possible experiments that are capable of supporting the view that long-range tropospheric propagation not associated with superrefraction is due to scattering by turbulent irregularities in the atmosphere.[1,2] Transmission by means of this scattering mechanism should involve a situation in which the energy arriving at the receiver is spread over a substantial angle, and by using sufficiently narrow beams this effect could presumably be demonstrated. Is it possible to demonstrate this effect with presently available equipment and technique?

Even at a wavelength as long as 10 cm there is now no serious difficulty in designing a paraboloidal antenna whose diameter D is 100 λ. It would be feasible to arrange an experiment using such antennas for both transmission and reception between two sites about 300 km apart. This is a distance such that, in the absence of superrefraction, the received power should be due to scattering by atmospheric irregularities and should therefore show a spread in angle of arrival. On the other hand, 300 km is a distance such that superrefraction should occasionally swamp the scattered signal, and the spread in angle of arrival should then be much less appreciable. It would be quite interesting to observe a change in the angle over which power is received as atmospheric conditions change from normal to superrefractive and back again. With such a path in mind, all calculations have been done for a distance of 300 km.

The transmitting and receiving antennas have been thought of as having identical circular apertures of diameter 100 λ, and parabolic aperture distributions with maximum illumination at the center and zero at the edges. This corresponds to a beamwidth of about 0.73 degree, a sidelobe ratio of about 25 db, and a power gain of about 48 db over an isotropic antenna. In most calculations, the antennas have been assumed to produce conical beams in free space of beamwidth 0.73 degree with no sidelobes, but in certain cases a somewhat more elaborate assumption has been made.

The wavelength plays little part in these calculations except in relation to antenna size. The results would apply to transmission over a 300-km path at almost any wavelength normally used in tropospheric propagation provided that antennas of beamwidth 0.73 degree are used. However, it is only at centimeter wavelengths that such antennas are, in fact, available.

It may perhaps be stated at the outset that the assumed beamwidths should be narrow enough to exhibit the scattering nature of the mechanism of transmission provided the experiment is properly carried out. It is important to realize, however, that to carry out the experiment properly, it is not sufficient merely to swing the receiving beam. This results in radical variation of the common volume formed by the cones corresponding to the two antennas. Indeed, the common volume can easily be destroyed if only one beam is swung. It is necessary to swing the transmitting and receiving antennas simultaneously so that the axes of the two beams continue to intersect. With narrow beams, this requires a fairly high degree of pointing accuracy. The calculations presented in this paper refer to synchronized beam swinging in which the axes of the transmitting and receiving antennas always intersect. So far as the authors are aware, previous attempts to observe the effects of atmospheric scattering by beam swinging have never employed synchronized beam swinging.

Section II describes the extent to which beamwidth must be reduced in order to demonstrate easily the

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[1] H. G. Booker and W. E. Gordon, "A theory of radio scattering in the troposphere," Proc. I.R.E., vol. 38, pp. 401–412; April, 1950.
[2] E. C. S. Megaw, "The scattering of electromagnetic waves by atmospheric turbulence," *Nature*, vol. 166, pp. 1100–1104; December, 1950.

angular spread in the scattered power. Following a general discussion of received power in Section III, Section IV considers the effect of antenna "gain" in which the concept of "aperture-to-medium coupling loss" is introduced. Section V deals with the variation of receiver power with angle in experiments involving synchronized beam swinging. Section VI deals with pulse distortion, and this is expressed in terms of the bandwidth of the circuit for communication purposes in Section VII, which includes a discussion of the effect of antenna beam angle on the ultimate signal-to-noise ratio of the circuit.

II. The Angle Over Which Received Power Is Spread

Following Booker and Gordon,[1] it is assumed that the power scattered by the atmosphere per unit solid angle, per unit incident power density, and per unit volume is

$$\sigma = \frac{\overline{\left(\frac{\Delta\epsilon}{\epsilon}\right)^2}(s/\lambda)^3}{\lambda\left[1+\left(\frac{2s}{\lambda}\sin\frac{\theta}{2}\right)^2\right]^2}, \tag{1}$$

where θ is the angle between the direction of incidence and the direction of scattering, $\overline{(\Delta\epsilon/\epsilon)^2}$ is the mean-square deviation of the dielectric constant of the atmosphere from mean, $s=2\pi$ times the scale l of the atmospheric irregularities, and λ is the wavelength. Exp. (1) is strictly appropriate to scattering in a plane perpendicular to the electric field of the incident wave, but for the small scattering angles in which we are interested the effect of polarization is negligible.

When Booker and Gordon[1] wrote their paper, no measurements had been made of $\overline{(\Delta\epsilon/\epsilon)^2}$ and l in the free atmosphere, so that they considered various possibilities, some of which are known now to be of no practical interest in tropospheric propagation. Since then, airborne refractometers[3,4] have come into use, and an analysis of a large number of observations by Crain, Straiton and von Rosenberg[5] has shown that values of l in free atmosphere range from about 20 to 130 meters, with a mean value of around 50 meters. There is no marked tendency for a variation with height except near the surface of the earth. Values of $\overline{(\Delta\epsilon/\epsilon)^2}$ range from 0.4×10^{-12} to 6.4×10^{-12}, with a significant tendency for a decrease with increase of height except close to the surface.

As pointed out by Gordon,[6] therefore, values of s in (1) are to be thought of as large compared with almost any wavelength normally considered in tropospheric propagation, and very large compared with the centimeter wavelength that would be involved in the 300-km experiment. Most of the scattering takes place in a beam within the angle λ/s radians of the forward direction,[1] and, at distances beyond the horizon, the receiver is not in the main scattering beam. Assuming that $\theta \ll 1$ radian, the basic scattering formula, therefore, reduces to

$$\sigma = \frac{1}{s}\overline{(\Delta\epsilon/\epsilon)^2}\cdot\frac{1}{\theta^4}\cdot \tag{2}$$

This formula applies only for scattering at angles such that

$$\lambda/s \ll \theta \ll 1, \tag{3}$$

but it is virtually certain that these conditions would be satisfied in the experiments under consideration.

It should be noticed that (2) for σ is independent of frequency and depends, not on $\overline{(\Delta\epsilon/\epsilon)^2}$ and s separately, but only upon the combination

$$\frac{1}{s}\overline{(\Delta\epsilon/\epsilon)^2} = S_P \tag{4}$$

which will be referred to as the "scattering parameter" of the atmosphere. The value of this scattering parameter is in the neighborhood of 10^{-14} to 10^{-15} meter^{-1} at the heights at which refractometer measurements have been made (up to about 12,000 feet). At the higher heights of interest in tropospheric scattering, the value may well drop to 10^{-16} meter^{-1}.

It has been shown by Gordon[6] that, under conditions when (2) is applicable, the horizontal angle over which scattered radiation arrives at a receiver is restricted to an angle whose magnitude is comparable with the angle of intersection of the tangent planes at transmitter and receiver. For the horizontal angle α_c in radians over which scattered radiation is spread at the receiver, Gordon gives the formula

$$\frac{2}{3}(d/R) = \frac{2}{3}\theta_0 \equiv \alpha_c, \tag{5}$$

where d is the distance between transmitter and receiver, R is the radius of the earth, modified to allow for normal atmospheric refraction, and $\theta_0 = d/R$. For the vertical angle over which scattered radiation arrives at the receiver, Gordon gives a similar formula assuming S_P varies inversely with the square of height above ground. Taking $d=300$ km and $R=8{,}000$ km, the angle of (5) evaluates to about 1.4 degrees. In these circumstances, the fact that the scattered signal is spread in direction of arrival will be quite noticeable for antenna beams appreciably less than about 1.4 degrees in width, but will not be an important feature for antenna beams appreciably wider than 1.4 degrees.

Practically all experiments made in connection with tropospheric scattering have used beam angles so wide that the effect of the spread in the direction of arrival of the scattered waves has been unimportant.[7,8] Thus Bul-

[3] G. Birnbaum, "A recording microwave refractometer," *Rev. Sci. Instr.*, vol. 21, pp. 169–178; February, 1950.

[4] C. M. Crain, "Apparatus for recording fluctuations in the refractive index of the atmosphere at 3.2 centimeters wavelength," *Rev. Sci. Instr.*, vol. 21, pp. 456–457; May, 1950.

[5] C. M. Crain, A. W. Straiton and C. E. von Rosenberg, "A statistical survey of atmospheric index-of-refraction variation," *Trans. I.R.E.*, vol. AP-1, pp. 43–46; October, 1953.

[6] W. E. Gordon, "Interpretation of diversity and fading measurements in tropospheric radio scattering," presented at the URSI Meeting, Washington, April, 1953 (to be published).

[7] K. Bullington, "Radio transmission beyond the horizon in the 40- to 4,000-mc band," Proc. I.R.E., vol. 41, pp. 132–135; January, 1953.

[8] G. R. Chambers, J. W. Herbstreit, and K. A. Norton, "Preliminary report on propagation measurements. From 92–1046 at Cheyenne Mountain, Colorado," Report No. 1826, National Bureau of Standards; 1952.

lington[7] reported that rotation of his antennas appeared to give roughly the polar diagram of the antenna, and that changes in the beam angle of the antenna gave the changes in received power that could be expected for rectilinear propagation. This was because he was using beam angles greater than α_c given by (5). In the suggested experiments, however, use could be made at both terminals of antennas having beamwidths of about 0.73 degree, and this is appreciably less than α_c which has a value about 1.4 degrees for the 300-km distance involved. The effect of the spread in the direction of the scattered wave should therefore be clearly visible in this experiment in a way that has not been possible in any previous observations.

III. Calculation of Power Received By Scattering

The formula given by Gordon[6] for power received vs distance was deduced on the assumption that the important scattering volume in the atmosphere is limited mainly by the scattering polar diagram of the atmosphere and this is appropriate when the beam angles of the antennas are larger than α_c. But, in the proposed experiment, the opposite would be true, thus the important scattering volume would be limited mainly by common volume of antenna beams as originally assumed by Booker and Gordon.[1] Taking free-space beam angles of antennas at both ends to be α, both horizontally and vertically, simple geometrical calculation shows that scattering volume is given approximately by

$$V = 2\left(\frac{d}{2}\frac{\alpha}{2}\right)^3 \frac{1}{\theta}, \tag{6}$$

and the mean scattering angle by

$$\theta = \frac{d}{R} + \frac{\alpha}{2} = \theta_0 + \frac{\alpha}{2}\cdot \tag{7}$$

In terms of the transmitted power P_T and the gain G of the antennas, the received power is given approximately by

$$P_R = \frac{P_T}{4\pi(d/2)^2}\cdot 2G\cdot \sigma V \cdot \frac{\lambda^2}{4\pi(d/2)^2}\cdot 2G, \tag{8}$$

assuming that the earth is a perfect reflector, and that $\int \sigma dV = \sigma V$. In free space, the received power would be

$$P_F = \frac{P_T}{4\pi d^2}\cdot G \cdot \frac{\lambda^2}{4\pi}\cdot G. \tag{9}$$

Hence the ratio of the power received to that which would be received in free space is

$$\frac{P_R}{P_F} = \frac{64\sigma V}{d^2}\cdot \tag{10}$$

From (2) and (6)

$$\sigma V = \frac{1}{32}\frac{d^3\alpha^3}{\theta^5} S_P, \tag{11}$$

$$= \frac{1}{32}\frac{d^3\alpha^3}{(\theta_0 + \alpha/2)^5} S_P, \tag{12}$$

on substituting for θ from (7). Using this value of σV in (10), we obtain finally for the ratio of the power received to that which would be received in free space

$$\frac{P_R}{P_F} = \frac{2d\alpha^3}{(\theta_0 + \alpha/2)^5} S_P. \tag{13}$$

This result is independent of wavelength for antennas of fixed beam angle but would depend on wavelength for antennas of fixed size. At distances such that

$$d \gg \frac{1}{2} R\alpha, \qquad (\theta_0 \gg \alpha/2), \tag{14}$$

(13) reduces to

$$\frac{P_R}{P_F} = \frac{2R^5\alpha^3}{d^4} S_P, \tag{15}$$

which corresponds to (5u) of Reference 1. The ratio of the power received to that which would be received in free space is thus inversely proportional to the fourth power of distance if S_P is independent of height. If, however, the scattering parameter of the atmosphere decreases with height, P_R/P_F decreases with distance faster than the fourth power.

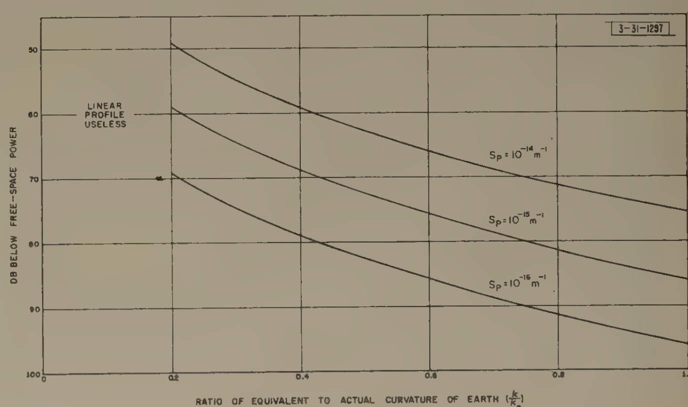

Fig. 1—Scattered power vs equivalent curvature of earth for a typical value and two extreme values of the scattering parameter of the atmosphere.

In the proposed experiment, we are interested in a fixed distance between transmitter and receiver, namely,

$$d = 300 \text{ km}, \tag{16}$$

and first in a fixed beam angle, namely,

$$\alpha = 1.27 \times 10^{-2} \text{ radian}. \tag{17}$$

In Fig. 1, above, three possible values of the scattering parameter have been assumed, namely,

$$S_P = \frac{1}{s}\overline{\left(\frac{\Delta\epsilon}{\epsilon}\right)^2} = 10^{-14}, 10^{-15}, 10^{-16} \text{ meter}^{-1}. \tag{18}$$

For these three values of the scattering parameter, and for the values of (16) and (17) for d and α, Fig. 1 shows P_R/P_F in decibels as a function of κ/κ_0 where κ_0 is the actual curvature of the earth (reciprocal of radius), and κ is the equivalent curvature of the earth taking into

account atmospheric refraction. Under normal conditions, the value of κ/κ_0 is about 0.8 but, under partially superrefraction conditions, a somewhat lower value may be appropriate. It will be seen from the figure that the power P_R received on the 300-km circuit should normally be in the vicinity of 82 db below the free-space value P_F. However, it might well be as low as 92 db under suitable atmospheric conditions, and could easily be as high as 72 db without involving any unusual refraction effects.

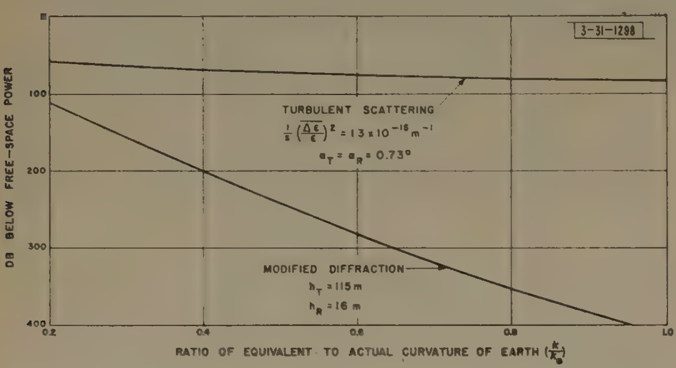

Fig. 2—Relation between scattering and modified diffraction fields. Distance = 300 km, wavelength = 8.3 cm, 100-λ apertures.

In the absence of scattering in the atmosphere, the power received over a 300-km circuit would be due to diffraction, modified somewhat by atmospheric refraction. A comparison between the two received signal powers is shown in Fig. 2. Under normal conditions, the received power due to diffraction should be more than 200 db below that due to atmospheric scattering. There is no doubt therefore that, under normal atmospheric conditions, the power received at a distance of 300 km will not be controlled by diffraction as modified to allow for normal atmospheric refraction. Under conditions of strong superrefraction, on the other hand, it is quite possible for the refracted signal to approach the value for propagation in free space and the refracted field would then exceed the scattered field. Most of the time, however, the field due to scattering by atmospheric irregularities should predominate.

It should be clearly understood that (13) applies only for beam angles α small compared with α_c, given by (5). If α is large compared with α_c, the effective scattering volume is not controlled by the beam angles of the antennas and (13) is, in accordance with Gordon,[6] replaced by

$$\frac{P_R}{P_F} = 0.86\frac{R^2}{d}S_P. \tag{19}$$

For $d = 300$ km, $R = 8{,}000$ km and $S_P = 10^{-16}$ meter^{-1}, (19) leads to a value of received power P_R which is 67 db below the free space value P_F.

IV. The Effect of Antenna "Gain"

It is interesting to study the variations of received power P_R with beam angle α as given by (15) and (19). This is illustrated in Fig. 3. Of course, in free space the variation of received power with beam-angle would simply be that normally described as arising from antenna gain, so that P_F is proportional to $(1/\alpha^4)$. It follows from (19), as already pointed out by Gordon,[6] that, if α is large compared with α_c [given by (5)], the power P_R received by scattering is also proportional to $(1/\alpha^4)$. This situation may be described by saying that, if beam angles are appreciably greater than α_c, the full gains of the antennas are obtained. If, however, α is small compared with α_c, (13) shows that P_R/P_F is proportional to α^3, with the result that P_R is proportional to $1/\alpha$. This is because, when α is less than α_c, the effective scattering volume in the atmosphere is proportional to α^3 and this largely outbalances the usual $(1/\alpha^4)$ variation due to normal antenna gain. Thus reduction of the common beam angle of transmitter and receiver down to the value α_c increases the power received in accordance with the improved gains of the antennas, but, when the beam angles are made less than α_c, the further improvement in power received becomes relatively minor.

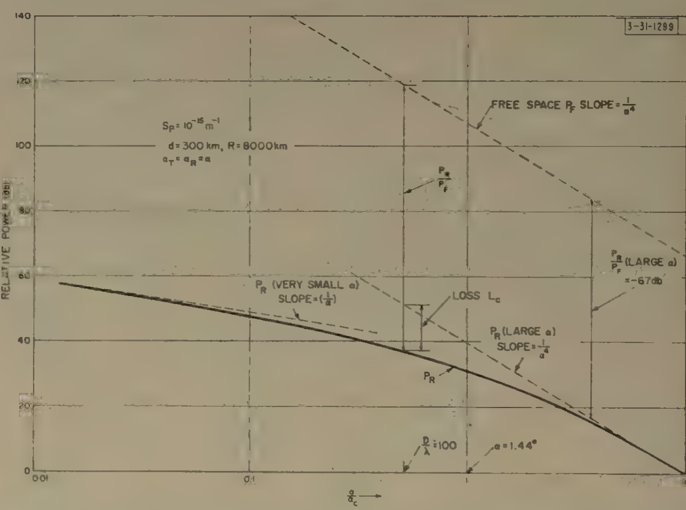

Fig. 3—Variation of received power P_R for small antenna beams with beam angle α. Values of P_R are shown relative to free space P_F. Scattering parameter S_p assumed to be constant with height above the ground.

The failure of the power received to continue proportional to $(1/\alpha^4)$ for $\alpha < \alpha_c$ is a phenomenon that is sometimes described as a "loss in antenna gain." However, we shall describe this phenomenon in terms of an "aperture-to-medium coupling loss," L_C shown in Fig. 4 (opposite). When α is less than α_c, the cone angles of the antennas are too small to encompass all the atmosphere capable of contributing to the scattered power at the location of the receiver. The antennas are not therefore making full use of the potentiality of the medium, and we shall say that there is an aperture-medium coupling loss L_C.

By dividing (19) by (15), we obtain the aperture-medium coupling loss L_C, for very small beam angles as a power ratio:

$$L_C = 0.43\left(\frac{d/R}{\alpha}\right)^3 = 0.43(\theta_0/\alpha)^3, \quad (\alpha \ll \alpha_c). \tag{20}$$

This formula applies for identical transmitting and receiving antennas $\alpha_T=\alpha_R=\alpha$. It further assumes that S_P is independent of height above ground. As α gets smaller, the height of the relevant scattering volume above ground decreases so that the appropriate value of S_P probably increases. Consequently, (20) gives values of loss that are probably too large.

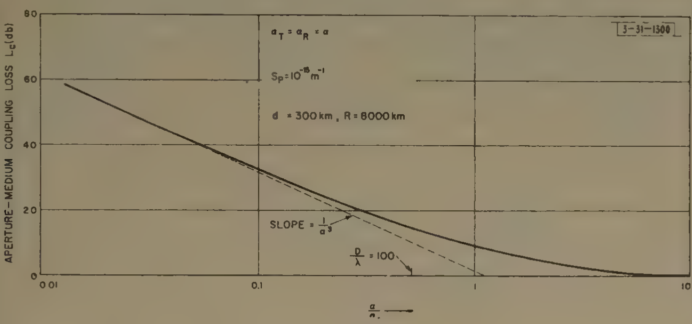

Fig. 4—Variation of aperture-medium coupling loss L_C for identical small-beam antennas with beam angle α. S_p assumed to be constant with height above the ground.

Using the solid curve for P_R and the dashed curve for P_R (large α) in Fig. 3, L_C may be determined, and is shown as the solid curve in Fig. 4. For large values of α, the "loss" is negligible and therefore independent of α; for very small beam angles ($\alpha \ll \alpha_c$), the loss varies as $(1/\alpha^3)$. For the 100-λ paraboloids contemplated on the 300-km path, an aperture-medium coupling loss of 14 db is obtained. This value is probably several decibels too large because the variation of S_P with height was neglected and also because of the rather crude assumptions made about the antennas.

In view of what has been said, it may be seriously questioned whether use of beamwidths smaller than about d/R is worth while. Purely from the viewpoint of power received, use of beamwidths less than about d/R is probably not worthwhile. However, for transmission due to scattering by atmospheric irregularities, beam angles also have an important influence on the bandwidth of the circuit for communication purposes (see Section VII).

V. Effect of Synchronized Beam-Swinging on Received Power[9]

When propagation is due primarily to scattering by irregularities in the atmosphere, the analysis of the effect of swinging the transmitting and/or receiving beams requires careful consideration. If the beam angles are large compared with the angle over which scattered radiation is spread ($\alpha \gg \alpha_c$), the effect of beam swinging is substantially the same as for rectilinear propagation, and the variation of received power is controlled mainly by the polar diagrams of the antennas. If, however, the beam angles are small compared with the spread due to scattering ($\alpha \ll \alpha_c$), quite a different situation arises. If one beam is swung and the other held fixed, the effect is somewhat complicated. But suppose that both beams are swung through equal angles in such a way that a common point of intersection of their axes is preserved. Then the variation of received power is controlled mainly by the scattering polar diagram of the atmosphere rather than by the polar diagrams of the antennas. Let us investigate the effect on received power of such synchronized swinging of the transmitting and receiving beams under conditions when the beam angles are small compared with the angle over which scattered radiation is important.

Let us assume that the angles of the transmitting and receiving antennas are equal, that the beams are initially directed horizontally in the vertical plane containing both antennas, and that each is then swung about a vertical axis through an angle ϕ on the same side of the great circle joining transmitter and receiver. The mean angle of scattering when $\phi=0$ is given by (7) but, if we assume that $\alpha \ll d/R$, this becomes

$$\theta = \frac{d}{R} = \theta_0. \qquad (21)$$

When the beams are swung horizontally to one side through an angle ϕ (small compared with one radian), the scattering angle becomes, by elementary geometry,

$$\theta' = 2\left[\left(\frac{d}{2R}\right)^2 + \phi^2\right]^{1/2}. \qquad (22)$$

As a result of swinging the beams, the scattering angle is therefore changed in the ratio

$$\frac{\theta'}{\theta} = \left[1 + \left(\frac{2R}{d}\right)^2 \phi^2\right]^{1/2}. \qquad (23)$$

From (2), the power scattered per unit volume is reduced correspondingly in the ratio

$$\left(\frac{\theta}{\theta'}\right)^4 = \left[1 + \left(\frac{2R}{d}\right)^2 \phi^2\right]^{-2}; \qquad (24)$$

and, from (6), the scattering volume is reduced in the ratio

$$\frac{\theta}{\theta'} = \left[1 + \left(\frac{2R}{d}\right)^2 \phi^2\right]^{-1/2}. \qquad (25)$$

The power received is therefore reduced in the ratio

$$\left(\frac{\theta}{\theta'}\right)^5 = \left[1 + \left(\frac{2R}{d}\right)^2 \phi^2\right]^{-5/2}$$

$$= \left[1 + \left(\frac{2\phi}{\theta_0}\right)^2\right]^{-5/2}. \qquad (26)$$

Expressed in decibels, this is plotted in Fig. 5 (following page) as a function of ϕ for $d=300$ km and for several values of the ratio of the actual to the equivalent curvature of the earth. The curve marked 0.8 may be regarded as representing normal atmospheric conditions.

[9] The authors are indebted to J. H. Chisholm of the Lincoln Laboratory, Mass. Inst. Tech., for many helpful discussions, particularly concerning this section.

It may be seen from Fig. 5 that, for the 300-km path under normal atmospheric conditions ($\kappa/\kappa_0=0.8$), the effect of turning the two beams to one side by 1 degree is to reduce the power received by about 7 db. If on the other hand we consider highly superrefractive conditions in which the power received is due predominantly to refraction instead of scattering, the effect of turning both beams aside 1 degree would be determined by the polar diagrams of the 100-λ apertures. The received power in this case would be reduced by more than 40 db. Thus the proposed 300-km link would provide the basis for an experiment demonstrating the existence of scattering due to atmospheric irregularities. The procedure would be as follows:

(1) Under highly superrefractive conditions, demonstrate that the effect of turning both antennas to one side about 1 degree is to reduce the power received by more than 40 db.

(2) Repeat the experiment under normal atmospheric conditions, demonstrating that the effect of turning both antennas to one side about 1 degree is at least 30 db less than under highly superrefractive conditions.

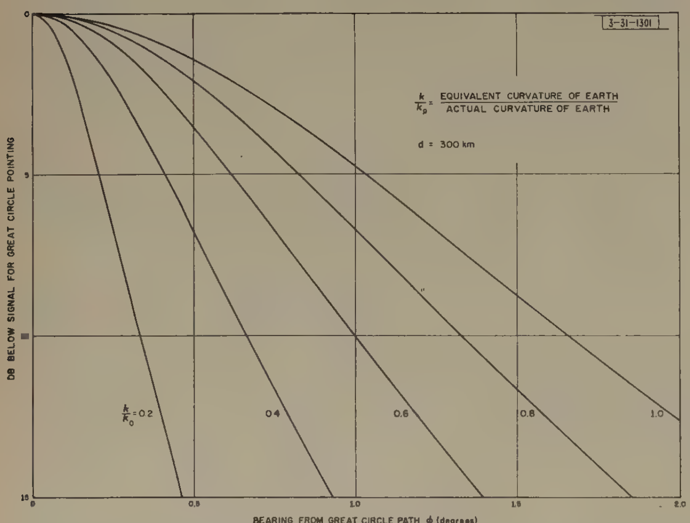

Fig. 5—Decrease in field due to turning T and R beams through equal azimuths. Indefinitely narrow beams are assumed, and a correction is required for beams of finite width (see Fig. 11).

A similar calculation to the above can be made if ϕ is kept zero but the axes of the two antenna beams are elevated by equal angles ψ. The ratio of the power received when the beams are elevated to that received when $\psi=0$ works out to be

$$\left(\frac{\theta}{\theta'}\right)^5 = \left(1+\frac{2R}{d}\psi\right)^{-5} = \left(1+\frac{2\psi}{\theta_0}\right)^{-5}, \qquad (27)$$

assuming that the scattering parameter of the atmosphere given by (4) is independent of height. In an atmosphere of scale height H, however, we might expect the scattering parameter to decrease with height h above ground, proportional to

$$\exp(-2h/H), \qquad (28)$$

and in this case the ratio of the power received for angles of elevation ψ, to that received when $\psi=0$, becomes

$$\frac{\exp[-\psi d/H]}{(1+2\psi/\theta_0)^5}. \qquad (29)$$

Figs. 6 and 7 show this power ratio, expressed in decibels, as a function of ψ, for $d=300$ km. In Fig. 6, it is assumed that $R=8{,}000$ km and curves are drawn for

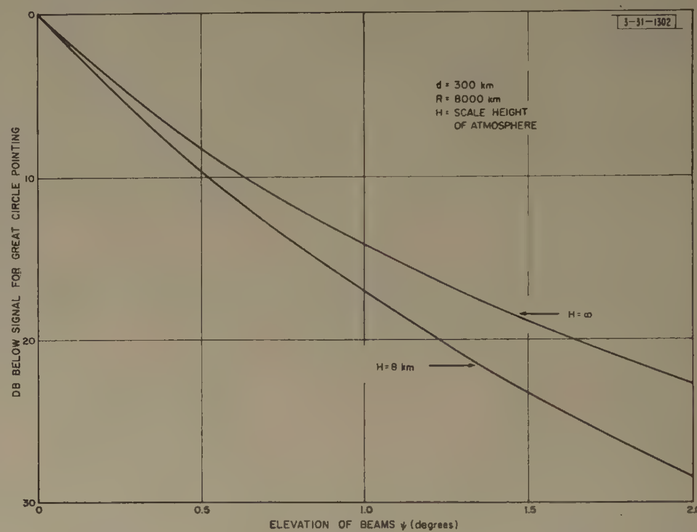

Fig. 6—Decrease in field due to elevating T and R beams equal angles, $R=8{,}000$ km. The calculation is done for uniform scattering with height, and for the expected exponential decrease with height. Subsequent calculations assume the latter.

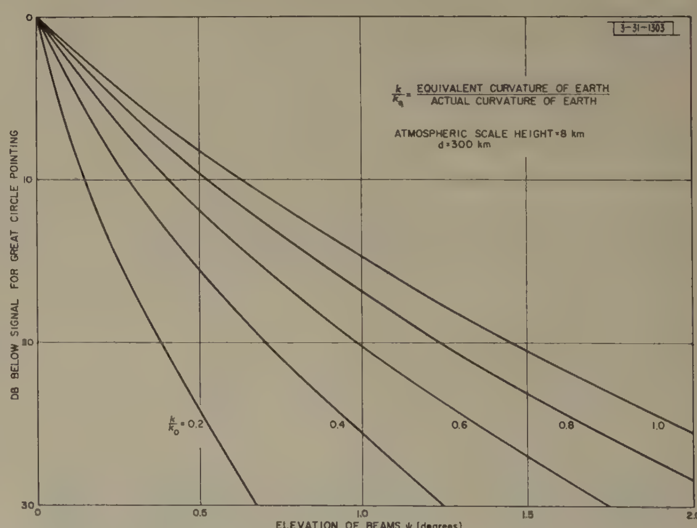

Fig. 7—Decrease in field due to elevating T and R beams equal angles, for several effective earth's curvatures. Indefinitely narrow beams are assumed, and a correction is required for beams of finite width (see Fig. 12).

$H=\infty$ and $H=8$ km. In Fig. 7, it is assumed that $H=8$ km, and various ratios of the equivalent to the actual curvature of the earth are assumed. It is clear that, in both the vertical and the horizontal directions, synchronized swinging of the beams on the 300-km circuit should produce appreciably smaller changes of re-

ceived power than would occur in free space, and this would be direct evidence that the power being received is spread over a cone larger than the antenna cones.

VI. Received Pulse Shape and Its Dependence on Beam Swinging

Different scattering elements within the scattering volume involve different delay times between transmission and reception, and this leads to some difference in shape between a transmitted pulse and the received pulse. Moreover, in the synchronized beam-swinging experiments contemplated in the preceding section, the mean transmission time will increase somewhat as the beams are swung away from the position of maximum signal. Let us make some calculations on these effects.

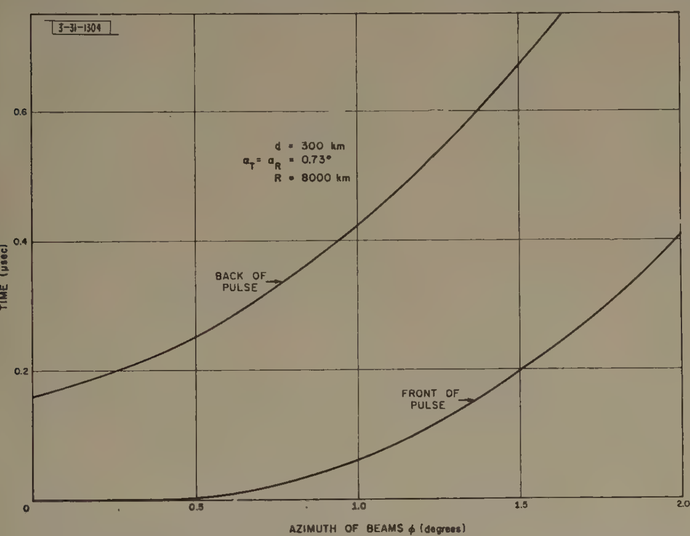

Fig. 8—Pulse delay and elongation due to turning T and R beams through equal azimuths.

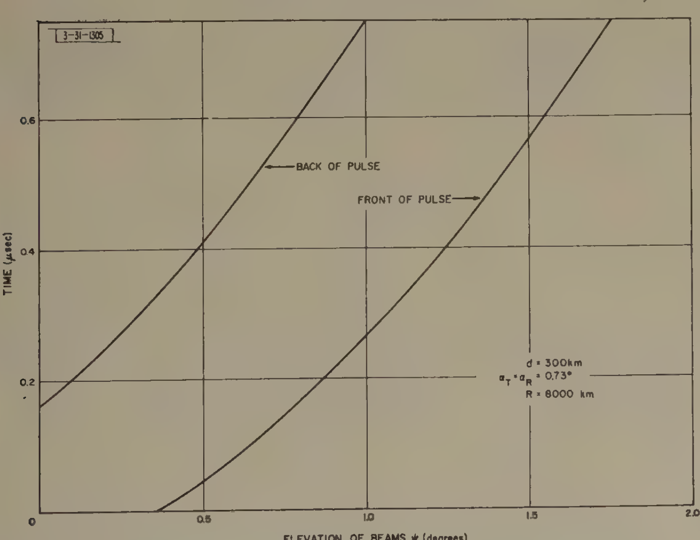

Fig. 9—Pulse delay and elongation due to turning T and R beams through equal angles of elevation.

Simple calculations may be made on the assumptions that the transmitted pulse is represented by an ideal impulse function and that the beams are cones of small semivertical angle $\alpha/2$. Within the scattering volume defined by the intersection of the cones, the points corresponding to minimum and maximum transmission time may be located. These define the front and back of the pulse arriving at the receiver. From such calculations, Figs. 8 and 9, in the previous column, were constructed.

When the azimuths and elevations of the beams are both zero, the received pulse at a distance of 300 km, corresponding to sending an ideal impulse over the path, is of length 0.16 μsec. Fig. 8 shows how the time of arrival of the front and back of the pulse vary as the two beams are turned to one side through equal angles. Fig. 9 gives the same information when the beams are elevated through equal angles at zero azimuth. It is clear from the figures that transmitted pulses of duration 0.1 μsec would be seriously distorted, but that there should be no marked distortion of pulses of duration 1.0 μsec.

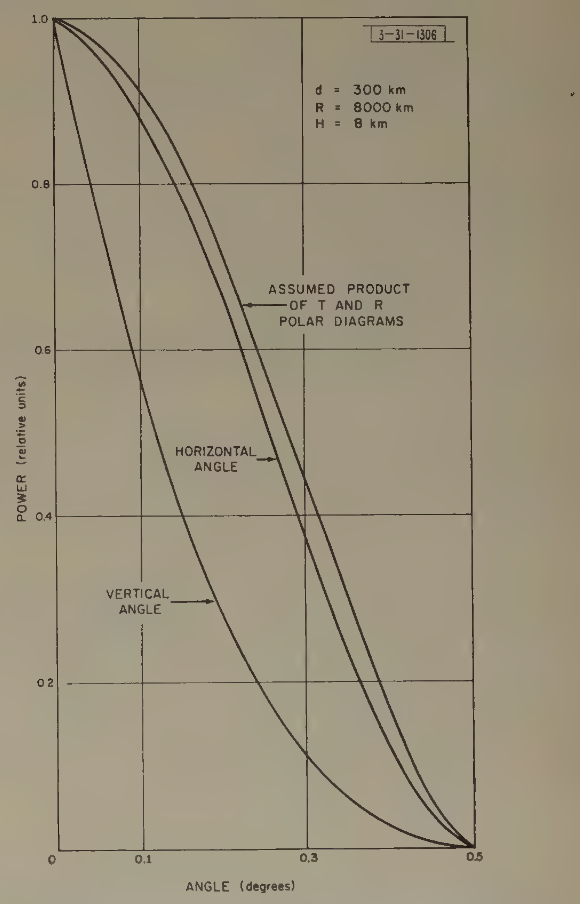

Fig. 10—Variation of response with horizontal and vertical angles of arrival, for great circle pointing. The assumed product of the T and R polar diagrams is shown. Each power polar diagram is assumed to have a parabolic main beam and no sidelobes.

For accurate calculations of received pulse shape, use should be made of actual polar diagrams of antennas instead of merely approximating to them by cones. In Figs. 10–14, on this and the following pages, use has been made, not of the complete polar diagrams of the antennas, but of a better approximation than a cone. The assumption is that the power polar diagram (power

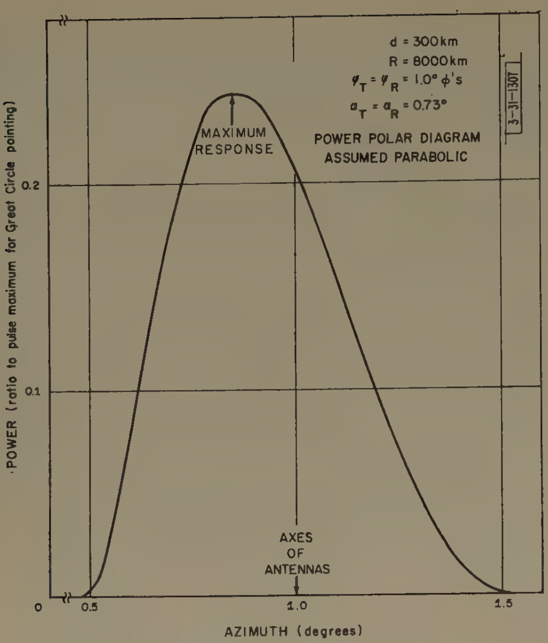

Fig. 11—Variation of response with azimuth when the axes of both antennas are at azimuth 1 degree.

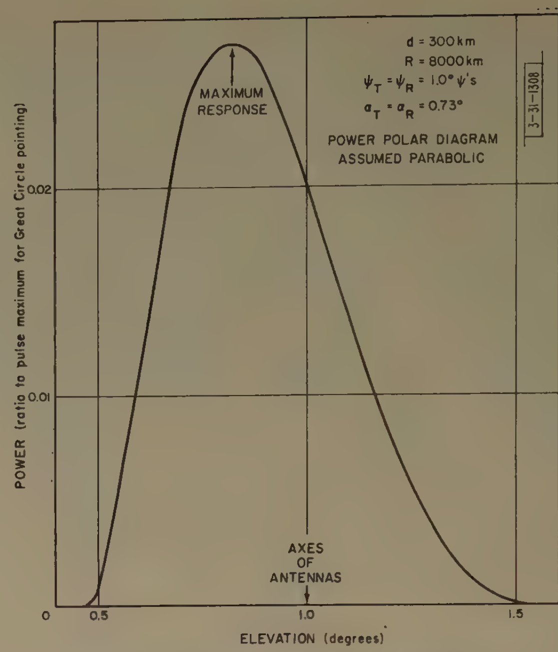

Fig. 12—Variation of response with angle of elevation when the axes of both antennas are at elevation 1 degree.

as a function of angle from axis) can be represented by a parabola and that sidelobes can be completely neglected. On this basis, the assumed product of the transmitting and receiving polar diagrams is represented by the upper curve in Fig. 10. Using this information, and the mean scattering polar diagram of the atmospheric irregularities derived from (2), we can calculate how the power received depends on direction of arrival and hence on delay time. This gives the shape of the received pulse corresponding to a transmitted pulse whose shape is an ideal impulse function, from which it is possible to deduce the corresponding information for any shape of transmitted pulse.

To simplify calculations, it has not been assumed that the antenna beams have circular symmetry about their axes. In each calculation, it has been assumed that the beam has a width (to half-power) of 0.73 degree in one plane, but is indefinitely narrow in the perpendicular plane. This assumption enables one to calculate separately the effect of horizontal and vertical multipath, but not of both simultaneously.

For the 300-km path with the axes of the beams at zero azimuth and elevation, the curve marked "horizontal angle" in Fig. 10 shows how the power received is distributed with horizontal angle of arrival. The curve marked "vertical angle" gives similar information for the distribution of received power with vertical angle of arrival. These curves are translated into received pulse shapes corresponding to a transmitted impulse function in Fig. 13 ($\phi_T=\phi_R=0$ degree) and Fig. 14 ($\psi_T=\psi_R=0$ degree). Fig. 11 refers to a situation in which the axes of the beams are turned to one side through 1 degree of azimuth. The curve shows the distribution of received power with horizontal angle. It shows in particular that maximum response occurs at an azimuth of 0.9 degree even though the axes of the beams are at azimuth 1.0 degree. This is because the parts of the beams nearer to the great circle path involve smaller scattering angles, and are consequently more effective. Fig. 11 is translated into a received pulse shape corresponding to a transmitted impulse function in Fig. 13 ($\phi_T=\phi_R=1.0$ degree). Fig. 12 refers to a situation in which the axes of the beams are elevated 1 degree at zero azimuth, and this curve is translated into received pulse shape in Fig. 14 ($\psi_T=\psi_R=1.0$ degree).

It is clear from all these calculations that a study of pulse distortion in scatter transmission over the 300-km circuit would require pulses with a rise time as short as 0.1 μsec.

VII. Radio-Frequency Bandwidth for Communication

The statements in the previous section concerning elongation of pulses in transmission due to atmospheric scattering may, of course, be translated into statements about the radio-frequency bandwidth of the circuit for communication purposes. In this connection, we consider only the situation in which the azimuths and elevations of the axes of the antenna beams are zero.

Consider a transmission path involving scattering at a point such that the azimuth at transmitter and receiver is ϕ, measured to the same side of the great circle path, and such that the angle of elevation at transmitter and receiver is ψ. A simple geometrical calculation shows that the corresponding path length exceeds that for $\phi=\psi=0$ by approximately

$$\Delta P = \frac{d}{2}\phi^2 + \frac{d}{2}\left(\frac{d}{2R}+\psi\right)^2 - \frac{d}{2}\left(\frac{d}{2R}\right)^2$$

$$= \frac{d}{2}(\phi^2 + \theta_0\psi + \psi^2). \qquad (30)$$

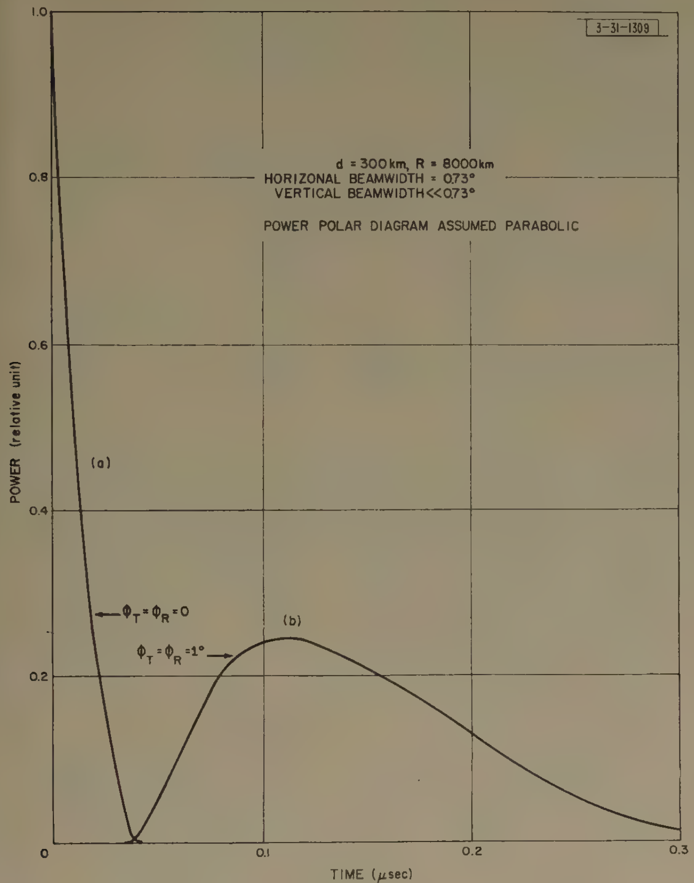

Fig. 13—Effect of horizontal multipath on the shape of the received pulses due to a transmitted impulse: (a) when azimuths are 0 degree, (b) when azimuths are 1 degree. Elevations are 0 degree.

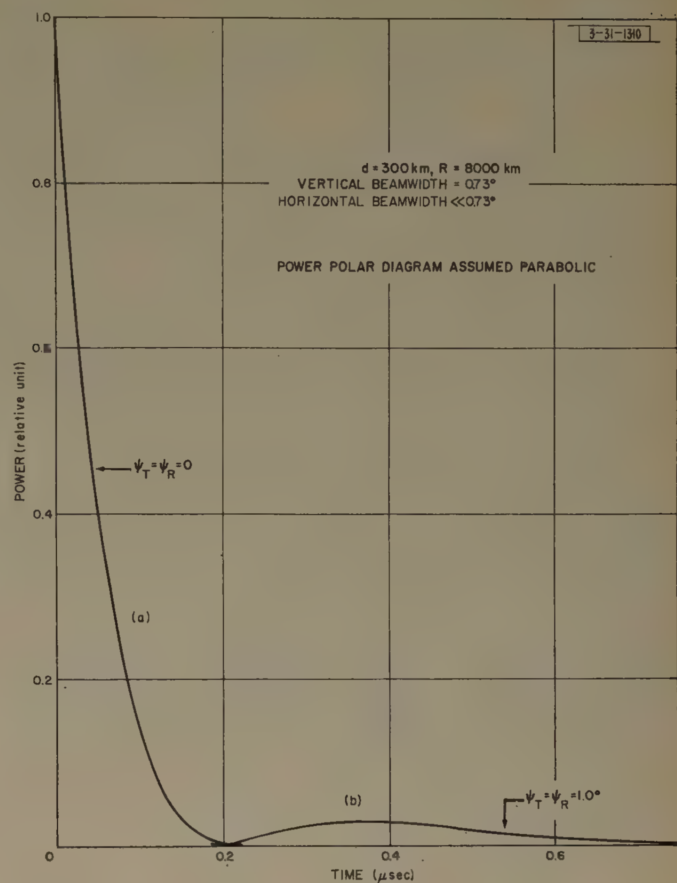

Fig. 14—Effect of vertical multipath on the shape of the received pulses due to a transmitted impulse: (a) when elevations are 0 degree, (b) when elevations are 1 degree. Azimuths are 0 degree.

The amount of pulse elongation involved in using beams of width α, less than α_c given by (5), is obtained approximately by putting

$$\phi = \alpha/2, \qquad \psi = \alpha/2 \tag{31}$$

in (30), giving

$$\Delta P = \frac{\alpha d^2}{4R}(1 + \alpha/\theta_0). \tag{32}$$

Since the assumption that α is small compared with α_c implies that α is small compared with θ_0, (32) reduces to

$$\Delta P = \frac{\alpha d^2}{4R} \qquad (\alpha \ll \alpha_c), \tag{33}$$

giving a bandwidth for communication purposes of

$$B = \frac{c}{\Delta P} = \frac{4cR}{\alpha d^2} \qquad (\alpha \ll \alpha_c), \tag{34}$$

where c is the velocity of light in free space. Eq. (34) shows that, for α small compared with α_c, the bandwidth of the circuit is inversely proportional to beamwidth. The narrower the beams, the less is the opportunity for multipath.

The bandwidth of a circuit for which the beam angles are greater than α_c has been considered by Gordon.[6] Gordon obtained a value of ΔP for the conditions $\phi = 0$, $\psi = 1/2\ \theta_0$. It might perhaps fit in better with (5) and (31) to put

$$\phi = \tfrac{1}{3}(d/R) = \theta_0/3, \qquad \psi = \tfrac{1}{4}(d/R) = \theta_0/4 \tag{35}$$

in (30), thereby slightly modifying the numerical coefficient in Gordon's formula. (These conditions correspond to the case where S_P varies inversely with the square of height above the ground.) The bandwidth for beamwidths large compared with α_c calculates to

$$B = \frac{c}{\Delta P} = \frac{4.7cR^2}{d^3} \qquad (\alpha \gg \alpha_c). \tag{36}$$

This bandwidth is independent of beam angle because, for beamwidths large compared with α_c, the degree of multipath is controlled by the properties of the atmospheric irregularities rather than by the antennas in use.

Variation of circuit bandwidth B with beam angle α is shown in Fig. 15 (next page) [10 log B (mc) vs α/α_c]. For large values of α ($\alpha \gg \alpha_c$), B, given by (36) is independent of α. For very small beam angles ($\alpha \ll \alpha_c$), B is given by (34) and varies inversely as α. For intermediate values of α, the curve is interpolated.

For $R = 8{,}000$ km and conditions (16) and (17), B from Fig. 15 is about 6.3 mcps which is about the reciprocal of 0.16 μsec derived from Figs. 8 and 9. For α

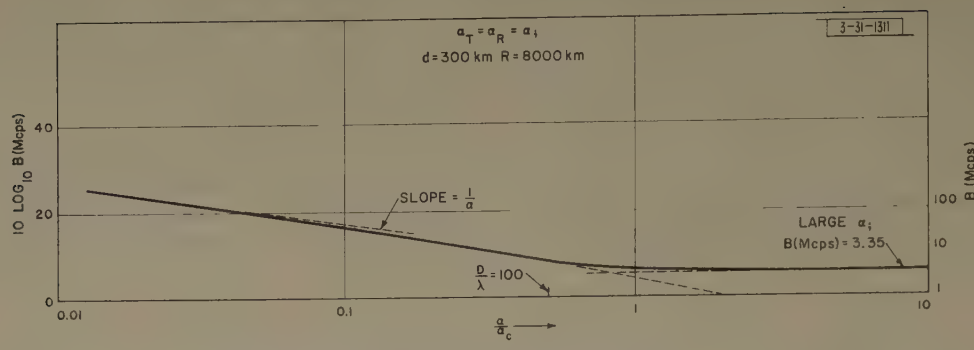

Fig. 15—Variation of transmission-circuit bandwidth B (in mc) for identical narrow-beam antennas with beam angle α.

larger than α_c, the circuit bandwidth, from (36) or Fig. 15, is about 3 mcps. Thus narrow beams, with α less than α_c, materially improve the capabilities of the transmission circuit insofar as bandwidth is concerned, even though the improvement in received power is minor.

The end result of varying beam angle on the signal-to-noise ratio S/N, is shown in Fig. 16, for conditions (16) and (17) and $R=8{,}000$ km. A typical receiver noise figure NF of 10 is assumed. It is further assumed that the transmitter power P_T and wavelength λ are such as to produce a signal-to-noise ratio of 10 db with $\alpha=10\alpha_c$. The variation of P_R is taken from Fig. 3. However, some care is necessary in discussing the variation of S/N with α. If circuit requirements are such that maximum possible bandwidth is demanded, then B is determined from Fig. 15 and the variation of S/N is determined in Fig. 16 from the curve labeled "bandwidth matched to transmission medium. . . ." In this case, narrowing the beam angle from large values results in a continued improvement in S/N until small values of α (small compared with α_c) are encountered, after which this ratio remains constant. Note that S/N does not decrease with decrease in α at values of α smaller than α_c.

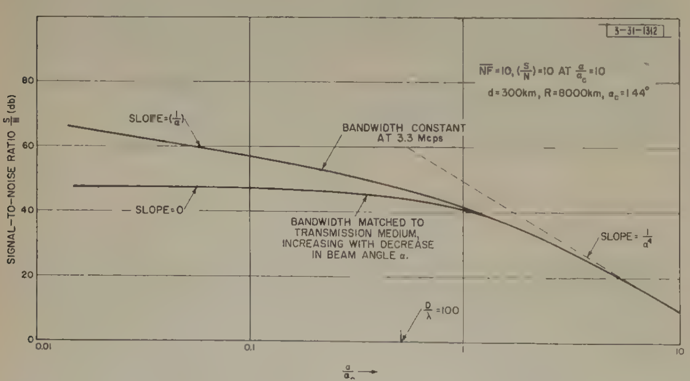

Fig. 16—Variation of signal-to-noise ratio for small-beam antennas with beam angle α.

If, however, a fixed bandwidth is to be used, equal to that given by the maximum allowed from (36)—i.e., about 3.3 mc in Fig. 16—variation of S/N with α is given by curve labeled "bandwidth constant at 3.3 mc."

Finally, if circuit requirements are such that only a moderate bandwidth is required, constant and less than (36), the variation of S/N with α is similar to the latter curve but shifted upwards by the reduction in noise with reduced bandwidth. Thus, if a bandwidth of 330 kc is satisfactory, the variation is S/N with α for the 300-km circuit will follow that given by the 3.3-mc curve but shifted upwards by 10 db.

VIII. Conclusions

Various transmission-circuit phenomena using identical narrow-beam transmitting and receiving antennas on paths extending beyond line-of-sight can be predicted approximately from the Booker-Gordon turbulent scattering theory.[1] Computations are based upon an exponential correlation coefficient and upon the assumption of off-beam scattering with scale of atmospheric irregularities large compared with wavelength.

It is shown that the behavior of such characteristics as signal level, circuit bandwidth, and pulse deterioration is dependent upon whether the antenna beam angles α are smaller or greater than a transitional value α_c. The value of α_c is given approximately by Gordon[6] as $2/3\ \theta_0$, where $\theta_0=d/R$, with d taken as the transmission distance and R as the equivalent earth's radius modified for refraction. For beam angles α much less than α_c, received power varies with α at a rate of $1/\alpha$ instead of $(1/\alpha^4)$ as with beams in free space, resulting in added loss interpreted as inefficient coupling between antennas and transmission medium. Again, for narrow beam angles ($\alpha \ll \alpha_c$), bandwidth capability of circuit varies as $1/\alpha$. Less multipath is involved with smaller beam angles. Simultaneous beam swinging, vertically or horizontally, should show variations differing from that for plane waves under normal tropospheric transmission conditions.

Calculations are given for a proposed 300-km path for 100-λ transmitting and receiving apertures, and for a scattering parameter assumed constant or, in some cases, decreasing exponentially with height. It is considered desirable to extend approximations made herein to a more exact description of the common scattering volume and to the variations of scattering parameter with height above ground. Finally, calculations should be made for other distances and for cases where transmitting and receiving beams are dissimilar.

Logical and Control Functions Performed with Magnetic Cores*

S. GUTERMAN†, ROBERT D. KODIS†, SENIOR MEMBER, IRE, AND S. RUHMAN†

Introduction

IN THE RELATIVELY short period since its first application to digital computers, the magnetic core having a square hysteresis loop, has firmly established its usefulness in high-speed large-scale storage[1-3] and for buffer storage[4,5] between input-output media, memory devices, and the heart or manipulative section of a computer. The magnetic core is particularly attractive because of its long life and consequent high order of reliability. A technique for performing logical functions with magnetic cores[6-9] will be described here, which is based upon the "single-line" shift register.[10] This method utilizes the inherent storage and amplification capabilities of the cores in order to make possible an all-core manipulative organ. Vacuum tubes are only used for the shift pulse sources which drive all the cores in common. Extensive tests proved the soundness of this technique.

"Single-Line" Shift Register

The main characteristics of the "single-line" shift register (see Fig. 1) will be reviewed briefly before proceeding with a discussion of the logical techniques based on it. The shift windings of all the cores are arranged in series to form what is called a "shift line." A source of current pulses, referred to as "shift" or "read-out" pulses, is applied to the shift line. Saturation in the direction of the field applied by the shift pulse is defined as the ZERO state of a core; saturation in the opposite direction is defined as the ONE state. Application of a shift pulse clears all the cores to state ZERO, thus generating a voltage in the output winding of any core which was previously in state ONE. This voltage charges the associated condenser through a diode; after the read-out, the condenser discharges through the succeeding core, thereby writing a ONE into it. Thus the information is stored

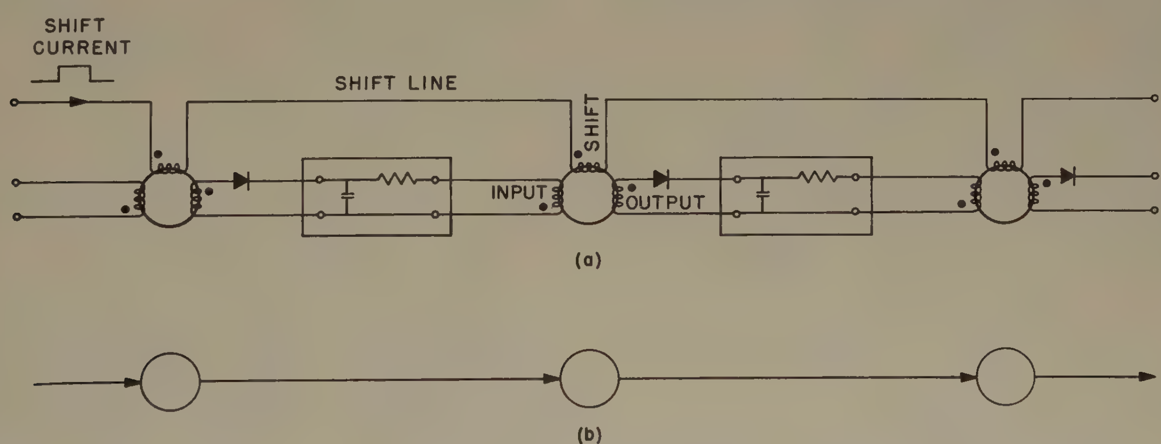

Fig. 1—Single-line shift register.

statically in the magnetic cores, and is advanced one stage with each application of the shift pulse, temporary storage being provided by the intercore link circuit. In Fig. 1(a), this link circuit is shown in its simplest form as a condenser-resistor combination.

Since many other types of temporary storage can be and have been used,[11] and the subject of this paper is not circuit design, the link henceforth simply will be represented by a four-terminal box. When the single-line register is applied to logical circuits, it becomes difficult to draw the circuit in detail. This has led to a shorthand notation for the register given in Fig. 1(b). It will be noted that the shift line is not shown and a common shift source for all the cores can be assumed unless otherwise indicated.

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Basic Functions

The three basic requirements necessary to make possible an all-core manipulative organ, such as the algebraic or control unit of a computer, are: (1) storage or delay, (2) amplification, and (3) two out of the three elementary logical operations, AND, OR, and INHIBIT (AND NOT). The storage or delay capabilities of the single-line register have already been described. The core provides indefinite static storage while, under dynamic operation, there is a delay of one pulse period per stage. Amplification is also inherent in the single-line register, as indicated by the fact that there is no attenuation in a register. In fact, each stage is a saturating magnetic amplifier, more energy being transferred to its output than is necessary to switch the core in the next stage. With proper design, it is possible to have this output switch more than one core. An amplifier stage, or branch circuit, whose output drives two other stages is shown in Fig. 2(a). The two driven cores are fed in series from the output of the amplifier stage. More than two stages can be driven from one, the upper limit being a function of the design. As a matter of fact, the authors have succeeded in driving five stages from one stage. Fig. 2(b) (above) gives the symbolic representation of the branch circuit.

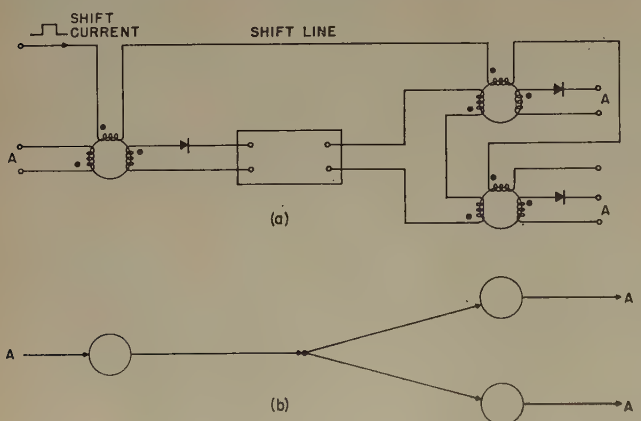

Fig. 2—Amplifier or branch circuit.

The two elementary logical operations from which the system under discussion is built up are OR and INHIBIT (AND NOT). Two ways of mixing (buffering) are used. The first, shown in Fig. 3(a), takes advantage of the diode normally associated with each core to buffer their outputs together into a common link circuit. The second, shown in Fig. 3(c), mixes the signals by feeding each into a separate input winding of the mixer core, the inputs being poled so that they tend to write a ONE in the core when they are energized. The first OR has the advantage that the buffered function is immediately available for use. On the other hand, the second OR has the advantage that the functions to be mixed as well as the buffered output are available for use, both OR circuits may handle more than two inputs, the maximum varying with the design. The symbol used for the two OR circuits, as shown in parts (b) and (d) of Fig. 3, indicates their nature and circuit arrangement.

The INHIBIT function [Fig. 4(a) (opposite)] is performed by a winding fed in opposite polarity to a normal input winding, and referred to as an INHIBIT winding. When this winding receives a ONE signal, it tends to cancel a signal appearing on the input winding. Since both windings are driven from similar stages, the two opposing signals are identical, within a small tolerance.

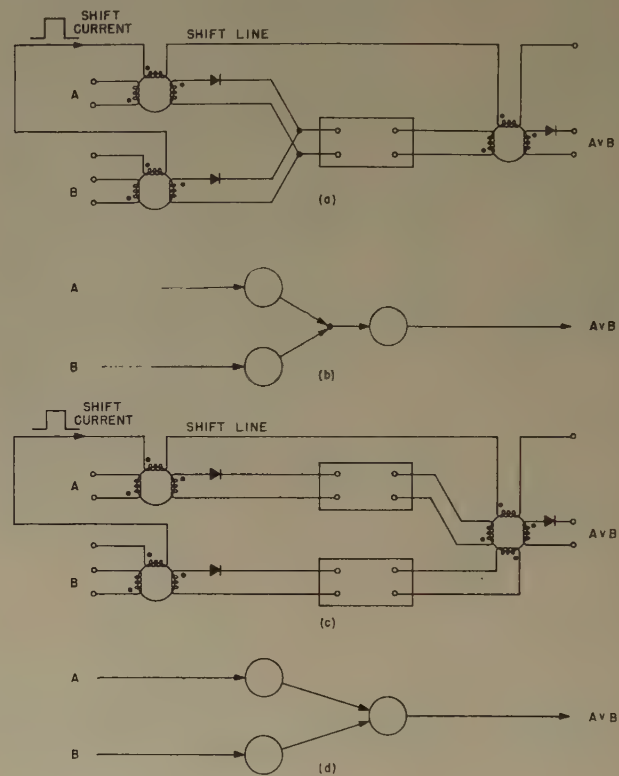

Fig. 3—Mixing or buffer circuits.

To insure that the inhibit signal will prevail, the inhibit winding may be made somewhat larger than the input winding. Where the input to be inhibited comes from a link circuit onto which the outputs of two or more cores were buffered, the inhibit winding is further increased to take care of the larger input appearing when all the buffered cores are energized simultaneously. Similarly, for a mixer core with two affirmative inputs, which is to be inhibited even when both inputs are energized together, the normal inhibit winding must be doubled. It should be pointed out that the inhibit winding places virtually no load on the driving stage, since no flux change takes place in the inhibited core. Hence, more than one INHIBIT may be driven from one stage, the upper limit again being a function of the design. Fig. 4(b) gives the symbol used for the inhibit circuit; the solid line across the path of information flow is meant to suggest the blocking or inhibit action. If it is desired to inhibit the input to a core by means of either of two signals, two inhibit windings may be used as in Fig. 4(c), or the two signals may be mixed in an OR which then drives the INHIBIT, Fig. 4(d).

These fundamental building blocks—the shift register, the BRANCH, the two ORs, and the INHIBIT—

form the whole basis for the all-core logical technique under discussion, all other logical functions being built out of them. The analysis and design of specific circuits will not be treated here; instead it will be shown how other elementary, as well as complex, circuits may be synthesized from these blocks and applied to computer systems.

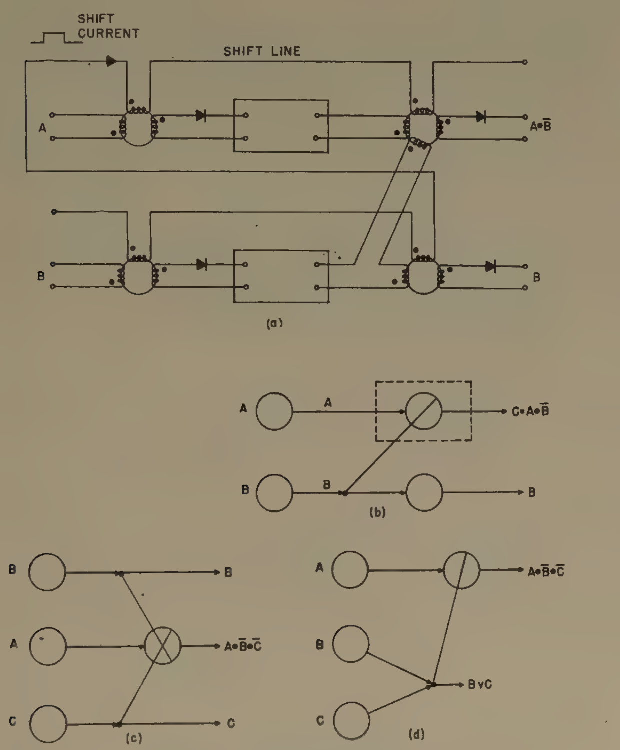

Fig. 4—Inhibit circuits.

Generation of Timing Pulses

Before proceeding with compound logical structures, it would seem desirable to discuss the generation of clock pulses, timing pulses, and other patterns or words for time-sequenced instructions and program control. A simple method, of course, is to connect a closed ring of shift register stages in which the desired information pattern is written initially and is then circulated by the shift pulses. This pattern is repeated at the output of each stage of the ring, but appears with a different timing (phase). This method has the advantage that all timing pulses are automatically available from a single ring, and any circulated pattern is available in every timing. The simplest application of this method is the generation of clock pulses (or continuous ONEs) from an initially set single register stage feeding back on itself.

A disadvantage of using a ring is that once an error occurs, the circulated pattern will remain incorrect until the shift pulse is turned off and the pattern rewritten. Another method of generating information patterns, which does not have this disadvantage, will now be discussed. The simplest example, a circuit which produces clock pulses (or continuous ONEs), will be described first. This circuit, called a ONE-Generator, is shown in Fig. 5(a). It consists of a core whose input winding receives a small dc current, poled so as to set the core to state ONE. The current is large enough to switch the core fully in the minimum period between shift pulses. Although this current actually opposes the shift pulse, it is normally so small by comparison that it may be neglected. If necessary, however, the number of turns in the shift winding of the core may be increased to compensate for this effect. The shorthand notation for the ONE-Generator is shown in Fig. 5(b).

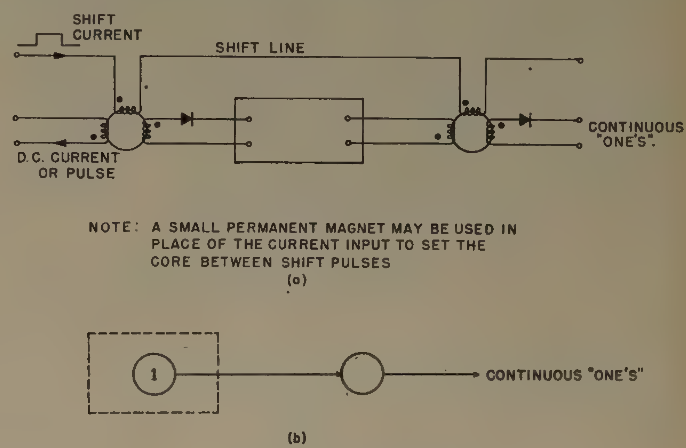

Fig. 5—ONE-Generator.

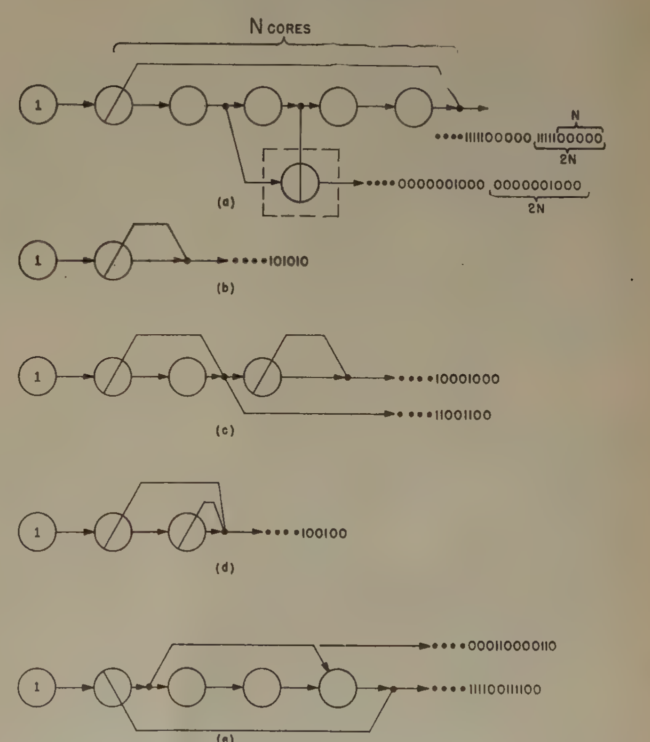

Fig. 6—Pattern generators.

More complicated patterns may be obtained by feeding a string of cores from a ONE-Generator and having feedback and advance feed between these cores that either inhibits or mixes with the main flow of information. In Fig. 6(a) such a pattern generator with an INHIBIT feedback from the last to the first core is shown.

If the number of cores in the line (aside from the ONE-Generator) is N, the resulting pattern is $2N$ digits long and consists of N ZEROs followed by N ONEs. For $N=1$ a frequency division by a factor of 2 results, as shown in Fig. 6(b). A circuit with $N=2$ followed by one with $N=1$ gives a frequency division by a factor of 4 [see Fig. 6(c)]. If in Fig. 6(a) a core external to the line is fed from one stage in the line and inhibited by the following or the preceding stage, the resulting pattern is a single ONE in a sequence of $2N$ bits, the phase of the ONE depending on the cores chosen. If the two cores are not adjacent but are two places away, the pattern will be two consecutive ONEs in a sequence of $2N$ bits. Two other interesting examples of pattern generation are shown in Figs. 6(d) and (e). Aside from their usefulness, the general theory of these circuits should be both interesting and challenging to the logician.

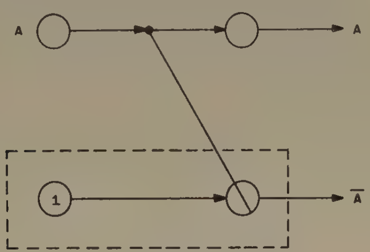

Fig. 7—Complementing circuit.

Synthesis of Other Elementary Functions

Proceeding now with the description of compound functions, the first one to be taken up will be a complementing circuit. In a system where OR and INHIBIT are the two fundamental logical functions, complementing is a very frequent and important operation. Referring to Fig. 7, the output of a ONE-Generator is fed into a core which is inhibited by A, giving ZEROs where there are ONEs in A, and vice versa.

$$\text{By symbolic algebra } 1 \cdot A = \overline{A}.$$

Thus, it requires two cores to produce from a binary information train A, its complement $\overline{A}$. With the use of the complement circuit, it is now easy to synthesize the important elementary function, the AND gate which is otherwise difficult to do reliably in a straightforward manner. Referring to Fig. 8(b), input B is complemented and $\overline{B}$ used to inhibit A, thus obtaining $A \cdot B$.

$$\text{By symbolic algebra } A \cdot \overline{(\overline{B})} = A \cdot B.$$

Another AND gate, not using a ONE-Generator, is shown in Fig. 8(a). Here B innibits A to give $A \cdot \overline{B}$, which in turn inhibits A, yielding $A \cdot B$.

$$A \cdot \overline{(A \cdot \overline{B})} = A \cdot B.$$

It should be noted that in the gate of Fig. 8(b), A need not be available until one pulse period before the result appears, while in the circuit of Fig. 8(a), A must be available two pulse periods ahead of the result, because of the need to form $A \cdot \overline{B}$. Obviously, applications in which $A \cdot B$ is to be fed back to form a new coincidence, $(A \cdot B)$ and B (next digit), one pulse period later, require the use of the Fig. 8(b) circuit.

A very simple function to synthesize with core logic is the exclusive OR, shown in Fig. 9. A inhibits B, and B inhibits A and the results are buffered together.

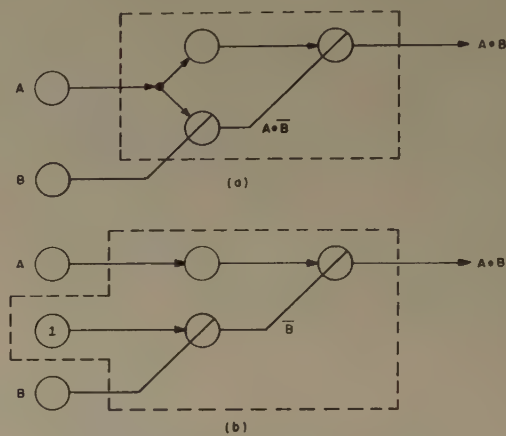

Fig. 8—Coincidence or AND circuits.

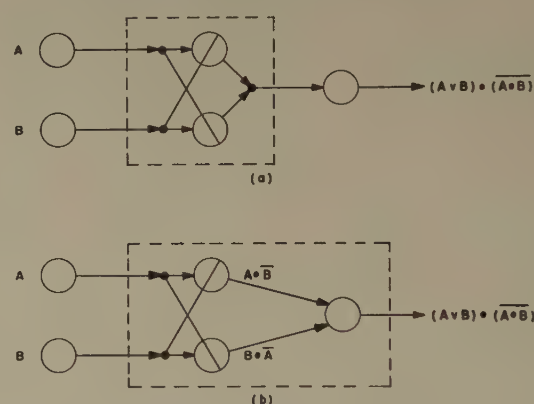

Fig. 9—Exclusive OR.

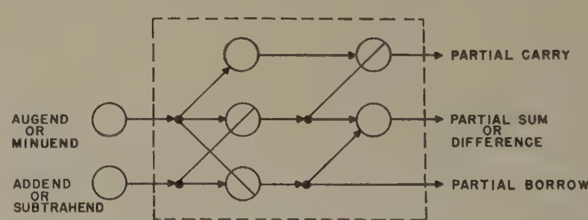

Fig. 10—Half-adder-subtracter.

$$\text{By symbolic algebra } (A \cdot \overline{B}) \vee (\overline{A} \cdot B) = (A \vee B) \cdot \overline{(A \cdot B)}.$$

Because of its simplicity, the exclusive OR is a very powerful tool in core logic. Its output is, of course, the partial sum or difference of the two incoming numbers A and B. The exclusive OR in Fig. 9(b) is in itself a partial subtracter, the partial borrow for $A-B$ appearing at the point marked $B \cdot \overline{A}$. By the addition of two cores to form a coincidence between A and B, as shown in Fig. 10, a half-adder as well as a half-subtracter is obtained, the partial carry being available at the output of the coincidence circuit.

With the single-line register, it is rather easy to build a flip-flop by the use of one stage feeding back on itself. As often as the contents of the register stage are read out, they are read back in again. To make provision for setting and resetting the flip-flop and for using its output, an OR, an INHIBIT and a BRANCH are used, all on the one-register stage. The set and feedback inputs may be mixed on the core as in Fig. 11(a), or they may be diode buffered into a common link network as in Fig. 11(b). Flip-flop (b) has the advantage that when it is set, its output is available one pulse time earlier than that of flip-flop (a). On the other hand, flip-flop (b) ties up the output of the setting core and always passes the set signal to the output, even when the reset is energized simultaneously. Two flip-flops interlocked so that the set for one is the reset for the other may be used to obtain two complementary outputs. Fig. 11(c) shows two flip-flops of type (b) arranged in this manner.

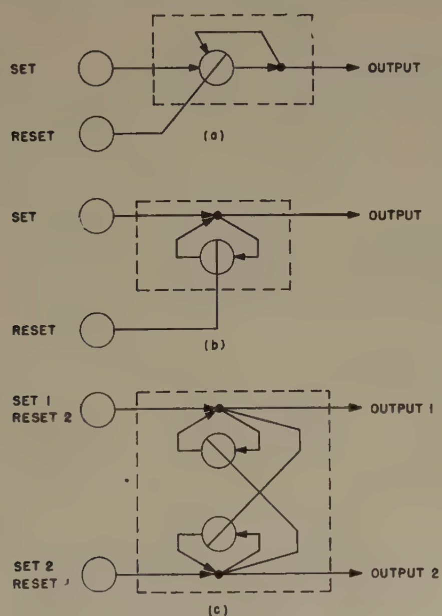

Fig. 11—Flip-flops.

Gating and Extracting

In the core logic, the magnetic core flip-flops can be used to control the flow of information throughout a computer as successfully as dynamic or static flip-flops in conventional computing technique. Here, however, anti-coincidence is mostly used for gating purposes. The output of the dynamic core flip-flop, shown in Fig. 11, when applied to an inhibiting winding of a core in a register, can stop information if a ONE is stored in it or let it pass if a ZERO is stored.

Two illustrations of the use of flip-flops to gate or control the flow of information are presented in Fig. 12. Part (a) shows a double flip-flop acting as a single-pole double-throw switch. The information to be switched branches into two cores, each inhibited by one of the two flip-flop outputs. Part (b) demonstrates how a flip-flop, used in conjunction with an exclusive OR, will gate out a train of information when reset, and complement this information when set. If, in the latter circuit [12(b)], the output of the core (P) is used as the input to set the flip-flop (F), the information which emerges will be the so-called 10's complement ($\bar{A}+1$). (Here one is added arithmetically to the least significant digit.) It is easy to see that the number starts to go through as a straight number until the first ONE comes along. This ONE will pass as a ONE. The remaining parts of the number will come out in complement form.

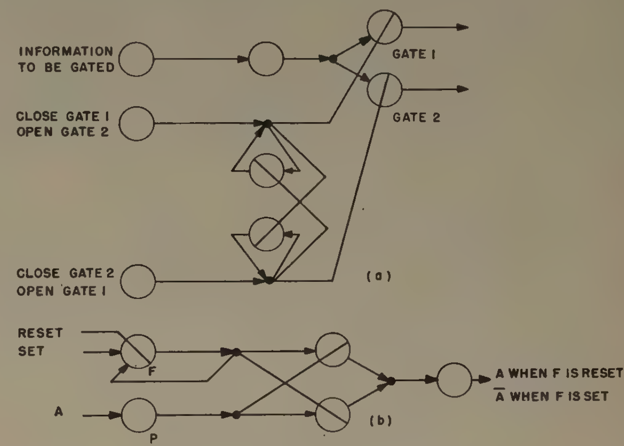

Fig. 12—Gating.

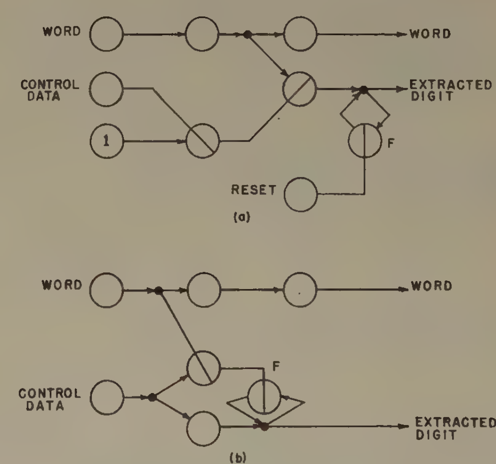

Fig. 13—Extracting and storing.

Another common computer and control function is that of extracting predetermined digits from a train of information. This may be accomplished by the conventional method of applying timing pulses as control data in coincidence with the digits to be extracted. An illustration of a circuit to extract a digit and also to store it for possible later use is shown in Fig. 13(a), along with a flip-flop, F, used to store the extracted digit. The ONE-Generator usually blocks the way in setting up flip-flop F (circuit a) except when a control data pulse appears and cuts off the ONE-Generator. F must be reset before each successive extraction. A similar circuit in which the flip-flop does not require an external reset is presented in Fig. 13(b).

Complex Logical Structures

A few illustrations of the use of the basic circuits, covered in the previous section, to build a complex logical structure will now be given. The extract circuit just described may be used to pass on information from a fast shift register to another slower one in arbitrary order (see Fig. 14). Two shift pulse sources, X and Y, are used such that Y is a subset of X. Every time a Y-pulse appears, a digit is transferred from the fast register to the slow register. Let N be the number of stages in the high-speed register, then a word circulating in the high-speed register will be transferred to the low-speed register in its normal order if there is one Y-pulse for every $[N+1]$ X-pulses, and it will be transferred in the reverse order if one Y-pulse occurs for every $[N-1]$ X-pulses.

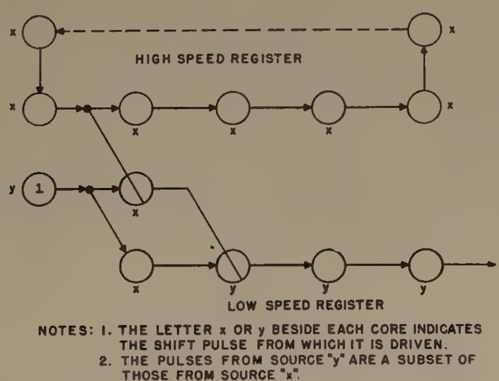

Fig. 14—Information transfer in arbitrary order from higher-to-lower-speed register.

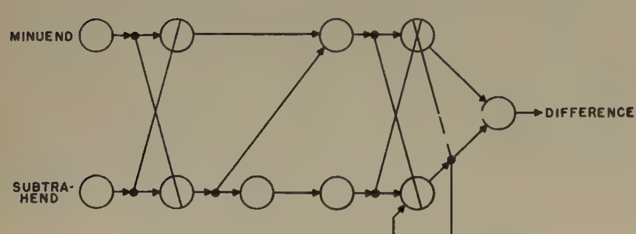

Fig. 15—Two-input subtracter.

It was explained before how a half-subtracter may be synthesized from three cores [see exclusive OR in Fig. 9(b)]. A two-input subtracter constructed from two half-subtracters is given in Fig. 15. Two additional cores are used to delay the partial borrow which appears one pulse time earlier than the partial difference, but in use should actually be one pulse time later, so as to act on the succeeding digit of the partial difference. A secondary borrow is fed back in the second half-subtracter and re-enters it one pulse time later, just as if it were a partial borrow. It can be shown that a partial and secondary borrow can never occur together. Thus, a two-input subtracter is obtained with eight cores and four pulse times of delay.

Similarly, a two-input adder may be constructed from two half-adders, as shown in Fig. 16. The first half-adder is of the type described earlier (see Fig. 10); the second is different and uses the AND gate of Fig. 8(b). The partial carry is delayed one place by an additional core and then applied to the AND gate, which generates secondary carries. If a secondary carry does result, it is fed back and acts exactly like a partial carry, one pulse time later. It would not be possible here to use the AND gate of Fig. 8(a), because a secondary carry will act two pulse periods later, which is too late for proper operation. Again it can be shown that a partial and secondary carry can never occur together. It should be noted that a two-input adder gives the same delay (four pulse times) as a two-input subtracter, but requires more cores.

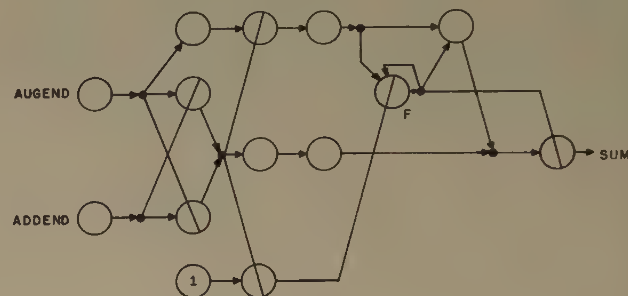

Fig. 16—Two-input adder.

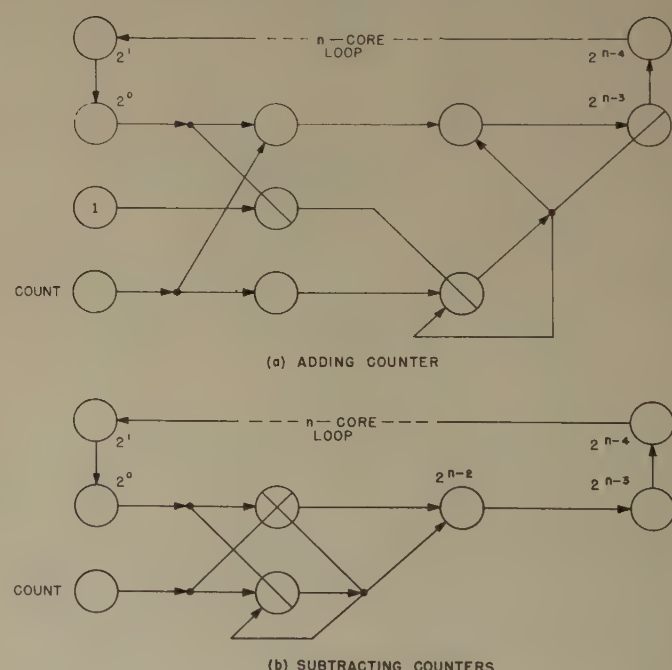

Fig. 17—Binary counters.

If it is desired to add or subtract just one digit from a number, the adder or subtracter discussed above may be simplified to the extent of removing the first half-adder or half-subtracter. Such an adder or subtracter may be used in conjunction with an N-core loop to form a counter of capacity 2^n, as shown in Fig. 17. The counter in part (a) uses the simplified adder. The register loop

stores the count, and is initially cleared to zero. When a count pulse appears, the loop is cycled completely and one is added to its contents. The counter in part (b) uses the simplified subtracter in a similar manner, except that all the cores in the ring are cleared to ONE, and each count pulse is subtracted from the contents of the loop, which now stores the *complement* of the true count. If necessary, the true count may be made available by the addition of a complementing circuit (two cores). In both of these counters the number in the loop must be recirculated to its initial position each time before a count can be admitted. Consequently, they are N times as slow as the core circuitry from which they are built. They are also limited in that they are unable to receive random counts. A more straightforward binary counter, which accepts random counts, has been devised using the same technique and will be described elsewhere.

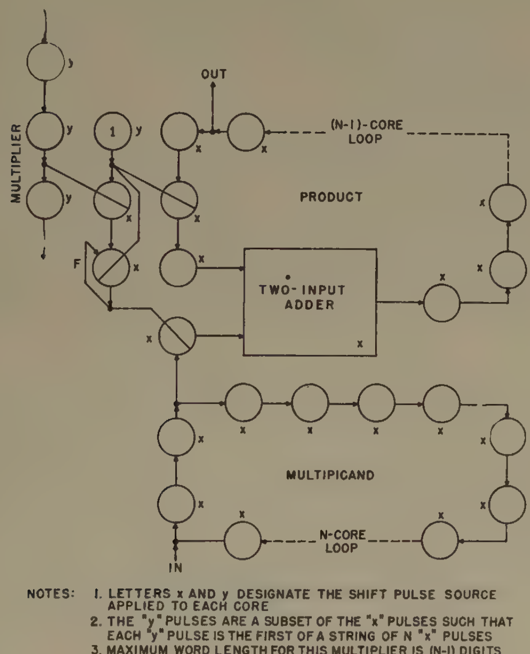

Fig. 18—Multiplier.

The full two-input adder (see Fig. 16) may be applied to build a multiplier as shown in Fig. 18. Multiplication is accomplished by the standard method of successive addition of the multiplicand to the partial product under the control of the multiplier. Two shift pulse sources are used, X and Y. The Y-pulses are a subset of the X-pulses such that every Y-pulse is the first of a group of N X-pulses. The multiplicand circulates in an N-stage loop and the partial product in an N-1 stage loop, including the delay in the adder. Thus, the partial product is advanced one pulse time after every addition. When the Y-pulse is applied to the ONE-Generator and the multiplier register, a digit is extracted from the multiplier and its complement is stored in flip-flop F by the next X-pulse. If the multiplier digit was a ONE, flip-flop F gates the multiplicand into the adder. If the digit was a ZERO, core F blocks the multiplicand so that it is not added to the partial product. At the beginning of each cycle, the least significant digit of the product, which has advanced "off-register" during the previous cycle, is eliminated and flip-flop F reset by the output of the ONE-Generator. The multiplier is capable of handling numbers up to $N-1$ digits long. The multiplicand can be fed into the loop at any point and the product can be taken out of the product loop at any point. The points arbitrarily chosen in the figure are such that the Y-pulse will coincide with the entering of the least significant digit.

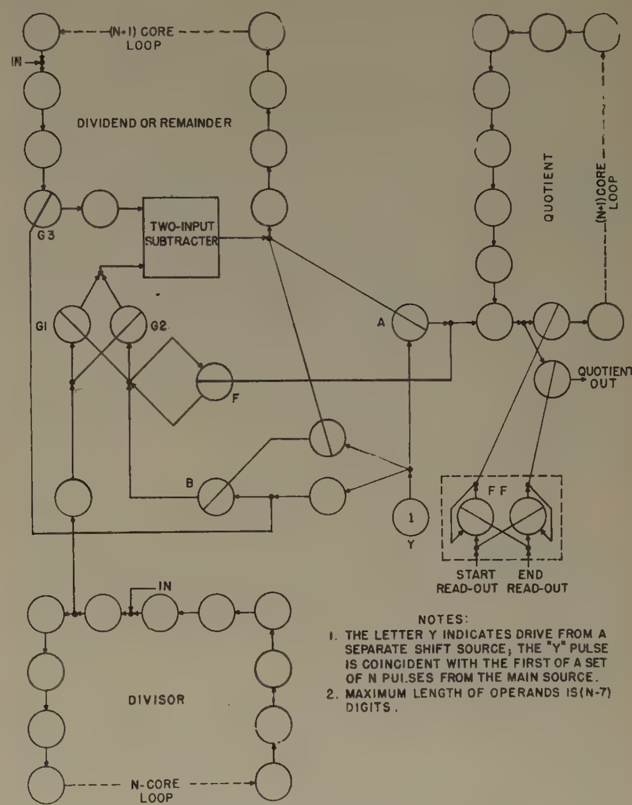

Fig. 19—Divider.

Similarly, the two-input subtracter (see Fig. 15) may be used to build a divider as depicted in Fig. 19. The division is accomplished by a process of successive subtraction of the divisor or its complement from the partial remainder, depending on the presence or absence of an over-all borrow. The complements of the over-all borrows form the quotient digits. The divisor, the dividend or partial remainder, and the quotient circulate, each in its own loop of such length that the quotient and the remainder are delayed one period with respect to the divisor after each subtraction. As in the multiplier, two shift pulse sources are used, X and Y, such that every Y-pulse is the first of a set of N X-pulses. Since the quotient digits are formed in order, starting with the most significant, the quotient is available in the proper form at the end of the division because of the one period of delay in the quotient loop.

The ONE-Generator is the only core shifted by Y-pulses. These pulses occur at the proper time for the

purpose of sensing and extracting the over-all borrows. The extracted value of the over-all borrow is stored in flip-flop F [Fig. 12(b)]. This flip-flop, together with gate cores $G1$ and $G2$, form the gating structure shown in Fig. 12(b), the function of which is to let either the divisor or its complement pass to the subtracter. The complements of the over-all borrow are fed to the quotient loop.

The output of the ONE-Generator also insures that after every subtraction the new lowest significance digit will be a zero by inhibiting it at core $G3$. Because it takes four pulse times to go through the subtracter and find out whether a final borrow will result, and it takes three more pulse times to set up gates $G1$ and $G2$, the maximum length of operands that can be handled is $N-7$ digits. However, the quotient may be obtained to any desired precision, provided the quotient loop is emptied whenever it has accumulated $N-2$ digits. The double flip-flop FF is used to gate out the contents of the quotient loop. As in multiplication, the points chosen to feed the divisor and the dividend into their loops are such that the least significant digits enter their loops coincidentally with the first Y shift pulse. Instead of using a ONE-Generator shifted by a special source, Y, it is possible to use a pattern generator similar to the one shown in Fig. 6(a). This will deliver a single ONE, followed by $(N-1)$ ZEROs.

Conclusions

All the basic circuits have been studied in some detail and many others have been built and tested at frequencies not exceeding 300 kc. This frequency is not in any sense an upper limit, but merely represents what has been done to date. Experimental results seem to indicate that the new logical technique is sound and highly practical. Work is in progress to apply the technique to a whole system in order to evaluate it more closely. The circuits presented in this paper are by no means the best, the most economical, or the only possible ones for solving these particular problems. The circuits shown serve only to illustrate the technique and show its possibilities, which are wide.

In conclusion, some of the advantages and limitations of the new all-core logical technique will be mentioned. The advantages are:

1. The magnetic core has an exceptionally long life.
2. No tubes are required for amplification and reshaping.
3. The power level may be made quite low in an all-core system.
4. A high order of reliability is attained because only a few tubes are needed to power the system with shift pulses.
5. In some applications this type of circuitry shows greater flexibility than existing methods.
6. The core circuits operate at very low impedance level, hence they are comparatively free from pick-up and cross-talk.
7. The problem of timing stability does not exist. In fact, the rate of the clock pulses may vary at random. The only condition is that they do not exceed the maximum rate permitted for a given design.

The limitations are:

1. The maximum frequency of reliable core circuitry is, at present, limited to under one megacycle.
2. This technique, being sequential in nature, is not very suitable for parallel operation.
3. In some applications, particularly where a host of inputs are to be gated or buffered, the new logical scheme is not as flexible as some existing methods.

Note

The Boolean algebra and its symbolic notation also apply to core logic with one condition, however, that the coincidence in time be respected. The best way is to assign a symbol for each shift pulse (T_1, T_2, T_3 etc.) and use these symbols together with others to form a logical equation. Referring to Fig. 4(b) for example,

$$T_2 \cdot C = (T_1 \cdot A) \cdot (T_1 \cdot \overline{B}) = T_1 \cdot A \cdot \overline{B}$$
$$T_3 C = T_2 \cdot A \cdot \overline{B}.$$

In this way, mistakes are avoided in analyzing a more complex structure.

Pin-Hole Camera Investigation of Electron Beams*

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Summary—**Electron beams produced by low- and moderate-perveance Pierce-type electron guns have been investigated by means of a pin-hole aperture. Data are presented showing the current density distribution and transverse velocity distribution of electrons at various points in the beam. The thermal velocity distribution is Maxwellian, with a width corresponding to the cathode temperature modified by the geometry of the flow. The transverse velocity distribution also reveals aberrations in the electron gun which has led to improved gun designs and a large improvement in the focusing of electron beams. The theoretical background for this work is given in a companion paper.[1]**

Introduction

IN SPITE OF the great progress in electron-beam type amplifiers (klystrons, traveling-wave tubes, etc.) in recent years, there is still much to be desired in the quality of high-density electron beams.

Most beam tubes now use a Pierce-type[2] of electron gun, which uses shaped electrodes to obtain a field distribution outside of the accelerated electron beam which matches the desired fields at the beam boundary. Several experimenters[3] have made measurements on the electron beam emerging from such guns and have determined that in a broad sense the gun is performing as planned. However, the fact that the focusing of electron beams requires much more than the design value of magnetic field, and gives less than expected transmission, leads one to suspect that all is not perfect in electron guns.

The possible defects in electron focusing and flow can be roughly divided into two categories. We will call them geometric and non-geometric causes. Pierce[4] has shown that for any given geometry- and space-charge-limited cathode emission, the electron trajectories are independent of voltage, provided the magnetic field is varied in accordance with $B \sim \sqrt{V}$. Thus we can differentiate between the two categories of defect. If the flow is controlled principally by the geometry (electrode configuration and applied fields, including space-charge effects), the beam trajectory will be independent of applied voltage (with appropriate scaling of the magnetic focusing field). Conversely, if we vary the applied potentials and magnetic fields in accordance with the above relation and get a large variation in transmission, we can be sure that there are important defects other than geometrical; such as thermal velocities, contact potentials, parasitic magnetic fields, or electron interaction or oscillations.

Experience shows a large change of electron flow as a function of voltage,[5] which means that there are certainly strong non-geometric defects in the focusing. The focusing at very high voltages is usually poor enough to lead one to believe that there also may be geometric defects to the flow.

Klemperer[6] made measurements of the flow through a slit in front of a rectangular cathode which revealed a spread due to thermal velocities and some anomalous effects which he attributed to electron oscillations. His observations reveal a powerful method of approach, but are inconclusive as far as our stated problem is concerned, because of the different geometry, and the fact that he did not account for the effect of beam spreading on the thermal-velocity distribution.

Beam Analysis by Pin-Hole Cameras

In order to obtain a better picture of the initial flow conditions, several tubes were made which used a conventional electron gun followed by a pin-hole aperture and a current detector (Fig. 1, following page). In some cases the measurements were made using a sealed-off tube [Fig. 1(a)] and a fluorescent screen to indicate current density and direction.[7] Alternatively a demountable, continuously-pumped structure as shown in Fig. 1(b) was used, and the current passing through the aperture was measured with a second movable aperture and a final collector. In either case we will refer to the plane of the current detection as the screen, and the current distribution at this plane as the image. By moving the first aperture through the beam one can measure the current density variation over the cross section. By also studying the image of the cathode in the detecting plane, one can determine a great deal about the initial transverse velocity distribution, and the flow geometry in the gun.

Ideally the flow in the gun would be laminar, and only the laminae passing through the aperture would reach the screen. This would excite a spot on the screen comparable to the size of the aperture with electrons from a similar size spot on the cathode. However, the existence of an initial thermal-velocity distribution permits electrons from a larger area to reach the aperture and create an inverted image of that area of the cathode on the fluorescent screen. The illumination of any point

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[2] J. R. Pierce, "Rectilinear flow in electron beams," *Jour. Appl. Phys.*, vol. 11, pp. 548–554; August, 1940.
[3] L. M. Field, K. Spangenberg, and R. Helm, "Cathode design procedures for electron beam tubes," *Elec. Commun.*, vol. 24, pp. 101–107; March, 1947.
[4] J. R. Pierce, "Theory and Design of Electron Beams," D. Van Nostrand Co., pp. 17, 19; 1949.
[5] Electron flow meaning actual beam trajectory, which may or may not be indicated by beam transmission in a given tube.
[6] O. Klemperer, "Influence of space charge on thermionic emission velocity," *Proc. Royal Soc. Lon.*, vol. 190, pp. 376–393; August, 1947.
[7] Under the very low current densities at the screen, this light output is proportional to current.

in the image is proportional to the number of electrons leaving a corresponding point on the cathode and having the direction necessary to reach the aperture. If there is a sufficiently wide range of transverse velocities, the whole cathode may be seen and the edge clearly defined.

Since the aperture as a lens has no definite focal length, the image might be taken to represent the character of the beam at any transverse plane between the cathode and the aperture plate. We can use the gun-design parameters and space-charge equations to compute the image magnification between any beam cross section and the screen[1] and if the flow at all resembles the ideal, we can interpret the image in terms of the conditions within the gun.

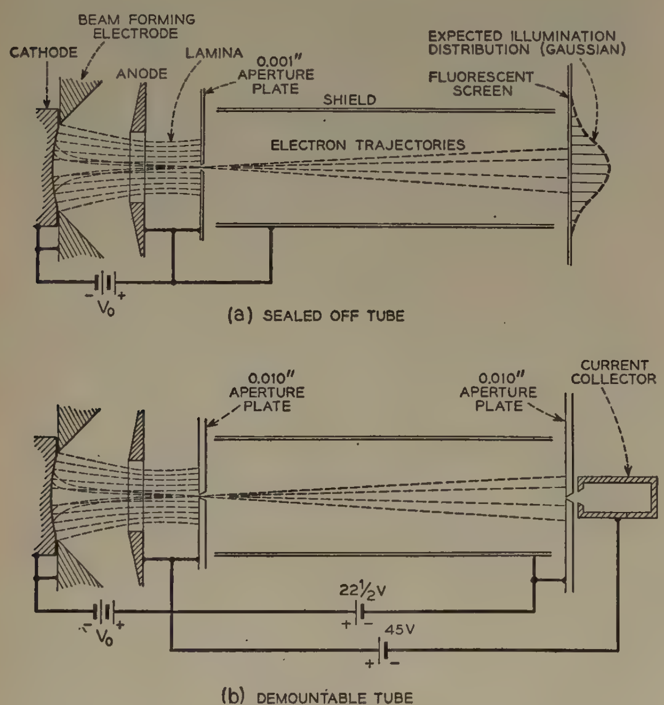

Fig. 1—Pin-hole camera tube. Case (a) uses a .001-inch diameter aperture and a fluorescent screen. Case (b) uses two .010-inch diameter aperture plates and a Faraday cage. By moving the aperture across the beam the density and lateral velocity distribution of electrons can be measured.

At the cathode the electrons might be expected to be distributed in velocity transverse to the direction of acceleration, according to the Maxwellian distribution.

$$dJ_c = \text{constant}\cdot e^{-mv_r^2/2kT}dv_c. \tag{1}$$

It is shown in the companion paper that the velocity spread should increase in proportion to the increase in proportion to the decrease in beam radius as the beam converges. Using this and the geometry of the tube, we have at the plane of the image

$$dJ_i = \text{constant}\cdot e^{-(e/kT)V_0M\,(r_i/s_i)}\,dri; \tag{2}$$

where V_0 = beam voltage (constant beyond gun anode), M = ratio of the beam radius at the aperture to the cathode radius. Other terms are defined at the end of the paper. This relation should apply fairly well out to the edge of the image (determined by the cathode edges).

Thus we should get a Gaussian distribution of illumination out to the edge of the image, and the width of the distribution curve should correspond to the effective temperature of emission modified by the beam-convergence factor M.

The transverse velocity distribution in the beam may be modified by nonuniform emission density and aberrations in the flow. However, by moving the aperture through the beam we can sample different parts of the beam cross section and determine the current from any point on the cathode to any point in the aperture plane. From this we can deduce the transverse velocity distribution at any chosen point on the cathode. If the distribution is Gaussian, we even can dare to use it as an indication of effective temperature of the emitted electrons. From observations of this kind we can get a measure of how closely the electron flow approximates the design intentions and perhaps get a clue as to what degrades the electron flow.

Using a pin-hole camera tube, then, we would expect to see an inverted image of the cathode with a Gaussian shape of illumination or current density, perhaps deformed by nonuniformities in cathode emission or aberrations in the focussing. When the pin-hole aperture is moved across the beam, the bright region should move a proportional amount in the image and maintain the Gaussian distribution of illumination. By measuring the characteristics and position of the image, we should be able to determine the following:

1. Beam size at the aperture.
   a. Density distribution over beam cross section.
   b. Beam convergence factor M.
2. Beam spreading at the aperture.
   a. Mean divergence from all causes.
   b. Laminar divergence (of most nearly laminar components).
   c. Divergence due to thermal velocities (effective beam temperature).
3. Transverse velocity distribution for electrons from any point in the cathode (translatable to effective emission temperature).
4. Beam density contribution of any point in the cathode to any point in the beam cross section at the aperture.

Departures from the predicted values should give a good measure of the gun's shortcomings.

Experimental Observations

Several convergent Pierce-type guns have been placed in the pin-hole camera structures described and the detailed focusing characteristics of the guns measured. In this paper two guns (Table I, page 301) will be used to illustrate the focusing characteristics observed.

Using the low perveance gun (No. 1), a series of photographs of the image on a fluorescent screen were

TABLE I

Gun No. 1 (Low perveance)	Gun No. 2 (medium perveance)
V =1,000 volts	V = 1,000 volts
I =3.0 ma	I =14.5 ma
r_c = .050 inch	r_c = .288 inch
r_{min}= .025 inch	r_{min}= .075 inch
θ =4 degrees 5 minutes	θ =13 degrees

taken for different positions of the pinhole aperture, together with measurements of image deflection and total image current. As the pinhole is moved, the image itself moves, due to the diverging character of the initial beam flow at the aperture plane, and at the same time current passing through the pinhole changes due to the non-uniformity of current density in the beam. The current density and beam spreading are plotted as a function of aperture position in Fig. 2. The interpretation of the beam spreading data is a little involved.

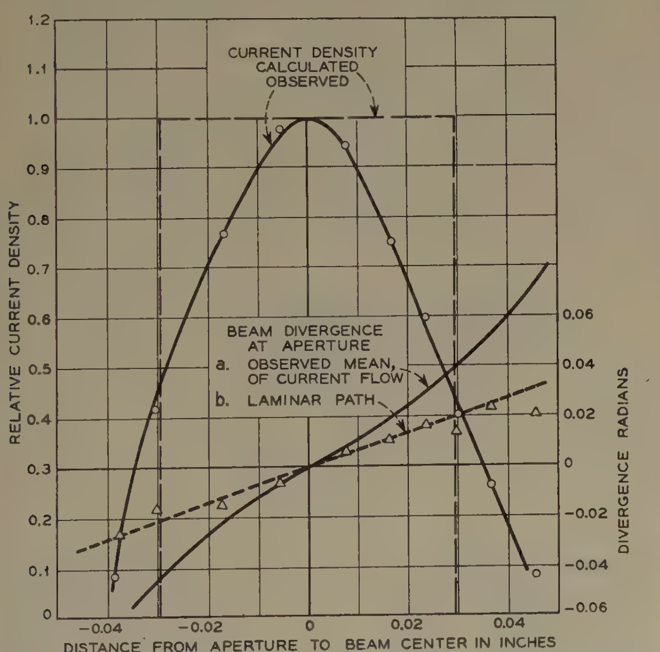

Fig. 2—Observation of current density and divergence in the beam of gun No. 1 (Table I). The points and solid lines give the measured data, the dashed lines are calculated for the design ideal.

If thermal velocities and other aberrations were negligibly small, the only point illuminated would be by laminar electrons and its position would be a good measure of the beam spreading. In presence of the other influences the laminar electrons are not so easily identified and we must trace them by other means. This can be done if the beam diameter is reasonably well defined at the aperture. A laminar electron will go from the aperture at r, ϕ to a point $(r/r_{max})R_{max}$, ϕ in the image, where r is the radial position of the aperture and R_{max} the image radius. Ideally this electron would come from a corresponding position $(r/r_{max})r_c$ in the cathode. To plot our data r_{max} was taken as the beam design radius at the aperture plane. The path so described is only known to be laminar at the plane of the aperture and does not necessarily follow a laminar path through the electron gun. Ideally a laminar electron would arrive at the peak of a gaussian image distribution. Curve b gives the calculated angle of flow for the gun and beam used, taking account of gun characteristics and space charge, while the triangular points show the measured spreading of the laminar electrons at the aperture plane which checks the predicted flow very well. The other curve of beam spreading is the angle of the mean direction of current flow through the aperture as a function of aperture position and is the deflection of the "center of gravity" of the image illumination. This is a better measure of the actual flow and in this case is much greater than the predicted value. The difference between curves a and b is the excessive spreading at the measured cross section and, by measurement of the image, can be separated into the effect of thermal velocities and aberrations. In the present case both are serious, as we shall see.

A curve of current density vs r^2 was integrated to obtain the increase in density over that at the cathode, and this used as a measure of the effective beam magnification. This is very close to the design value (M=.59). Obviously, however, the density is far from uniform over the cross section, and the beam diameter is not sharply defined. Projecting the observed flow backwards we see that the mean convergence within the gun must be greater than the design value.

The image magnification for this tube (i.e., the ratio of the spot diameter to the cathode diameter) is almost exactly that calculated from the ideal case. The image of the cathode is strongly illuminated completely to the edge, however, and therefore is limited by the cathode size rather than by transverse velocities, as can be seen in Fig. 3(c). When the aperture is moved, the illumina-

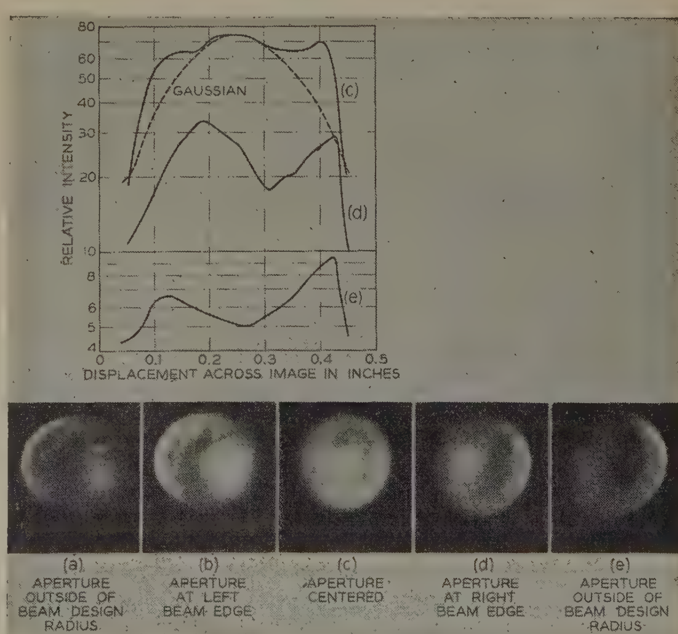

Fig. 3—Fluorescent screen images obtained from gun No. 1 in a pinhole camera tube. The curves give the measured illumination of the corresponding image. Compare this with Fig. 7.

tion varies quite differently from that expected (see previous section). The edge of the image toward which the aperture is moved remains bright, whereas with proper flow it should become more dim. Also a bright region moves from the center toward the opposite edge of the image in roughly the proportion that the aperture is moved through the beam. When the aperture is outside of the nominal beam diameter, there is no evidence of the latter bright region, but the cathode edge (the wrong one) is still clearly defined. It is clear that the spread in transverse velocities is much greater than the Maxwellian distribution of (1).

If we assume the screen luminescence to be proportional to the current density of the impinging beam, we can get a quantitative measure of the current distribution in the images. Several photographs of the cathode image were taken with different exposures and were used to calibrate film density vs exposure. Densitometer measurements of the films give the light intensity distribution over the image as indicated in Fig. 3(c).

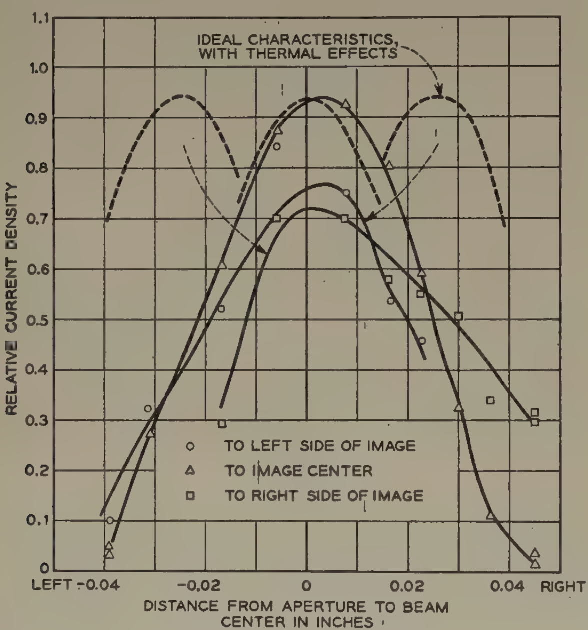

Fig. 4—Current density at given points in the image as a function of aperture position. An ideal gun, allowing for thermal effects, should give the dashed curves. The measured curves reveal excess beam convergence.

We can study the transverse velocity distribution in two ways. One is to measure the distribution of image illumination as in Fig. 3(c), and the other is to measure the illumination of a point in the image as a function of aperture position. Fig. 4 shows the current density at three points in the image as a function of aperture position. The current at one side of the image comes from the opposite side of the cathode, indicating a severe crossover in the electron flow.

It is helpful to plot current density vs transverse electron velocity at the aperture, for electrons from a fixed part of the cathode. We use as a zero velocity reference the position in the image corresponding to the point which would be illuminated if we had the design conditions for ideal laminar flow. This point moves across the image in proportion to the motion of the aperture through the beam, while the current to the reference point in the image varies. Plotting the log of this current vs transverse velocity squared, thermal velocities would give a straight line, as in (2).

$$\text{Log } J_i = \text{const}\left(\frac{e}{kT} V_0 M^2 \frac{r_i^2}{z_i^2}\right). \tag{3}$$

shown as a dashed curve in Fig. 5. The points taken for the center of the image correspond closely to the theoretical curve, indicating that the transverse velocity distribution (or electron temperature) is near the calculated value.

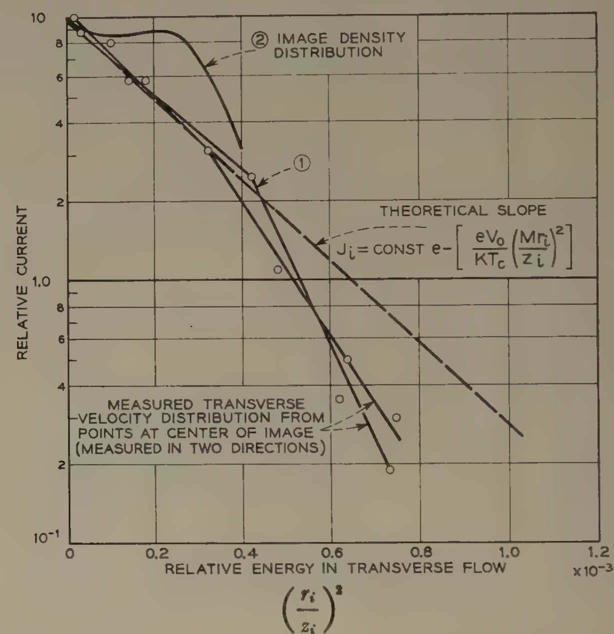

Fig. 5—Transverse velocity distribution. In Curve 1 is plotted the current arriving at the center of the image as a function of the aperture position (measured by laminar electron displacement in the image). Curve 2 is the distribution of image current density, and the dashed curve is the theoretical Maxwellian distribution which is predicted for either measurement for an ideal gun. Curve 2 reveals strong aberrations in the gun.

If we plot the curve of Fig. 3(c) on these coordinates, we get quite a different curve (curve 2 of Fig. 5) instead of the straight line velocity distribution, showing the serious effect of the gun aberrations.

The measurements described apply strictly only to the condition of the beam at the aperture. Of course the important question is: What does all this mean regarding the focusing conditions within the gun? Since the general flow conditions and image magnification are close to the calculated values, we feel reasonably safe in projecting electron paths backward through the pinhole to determine corresponding conditions at the cathode; and thus, we can be rather certain of the source of any particular nonuniformity. For instance, the bright edge (or ring) in the images of Fig. 3, if projected back to the anode aperture is still well within the

maximum beam radius. Therefore, it is not likely that it is caused by aberrations in this region. If we assume that the transverse velocity compared to any given lamina of the design ideal is constant, it projects almost exactly to the cathode edge. For this reason, we believe that the observed image does give a picture of conditions near the cathode. Thus Fig. 5, taking into account the convergence factor M, indicates a normal thermal velocity distribution at the cathode center and an excess velocity directed toward the beam axis of up to 0.6 volt near the edge of the cathode.

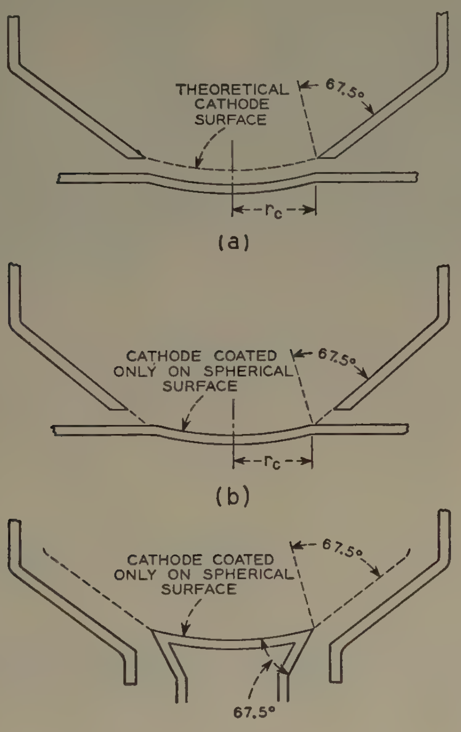

Fig. 6—Various beam-forming electrode designs providing for thermal and electrical insulation of the cathode. (a) Cathode is retracted from design position. (b) Beam-forming electrode is shaved off to clear the cathode. (c) A negative electrode is used to bring the cathode potential equipotential to the proper position.

In this gun the cathode was retracted from its design position by .002 inch to provide the necessary heat insulation from the beam-forming electrode [Fig. 6(a)]. Since voltage in the gun varies approximately as

$$V = \frac{V_0}{(z_a)^{4/3}} (z)^{4/3} = 4.8 \cdot 10^3 z^{4/3},$$

at 1,000 volts anode potential, the displacement of the beam forming electrode is equivalent to an error of about 1.2 volts. Thus it seems probable that the excess transverse velocities are caused by a mismatch of the boundary conditions at the cathode edge. This hypothesis is further verified by biasing the beam-forming electrode. A negative bias increases the edge intensity described and a positive bias reduces it, but does not correct it, for a sharp hair line remains, presumably caused by reflection of electrons from the edge of the beam-forming electrode.

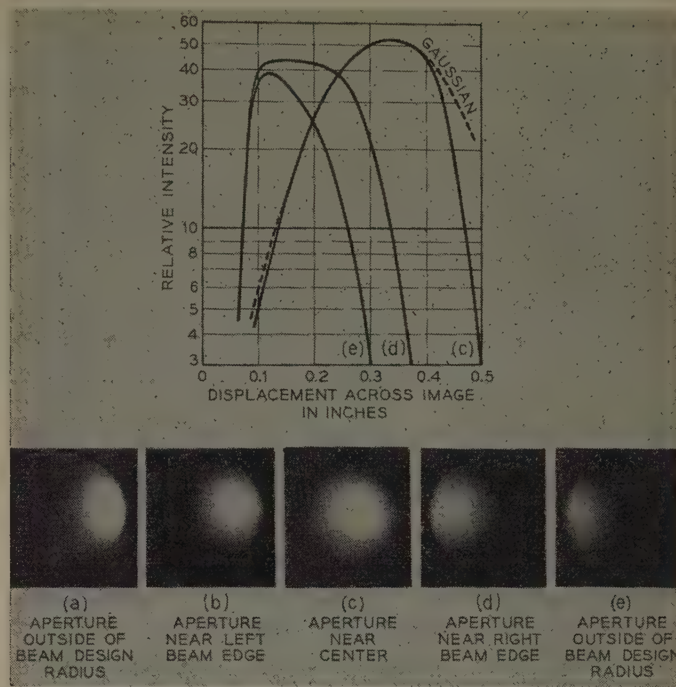

Fig. 7—Photographs of images from a gun having the same electrical design as gun No. 1, but using the negative beam-forming electrode design of Fig. 6(c). These pictures reveal no significant departure from the ideal.

Improved Electron Gun Design

There are several ways one can approximate the cathode boundary conditions required in a Pierce-type electron gun. Ideally there should be a cathode potential surface extending away from the cathode edge making an angle of $67\frac{1}{2}$ degrees with the outside of the beam. In practice it is desirable to insulate this surface thermally and electrically from the cathode, and this is commonly done by retracting the hot cathode from a cathode potential electrode as shown in Fig. 6(a). An alternative is to remove part of the electrode as in Fig. 6(b), limiting the emitting material to a smaller diameter than the hole in the electrode. Another way is to use a specially shaped cathode and a negatively biased electrode as in Fig. 6(c).

In order to minimize the mismatch in the boundary conditions at the cathode edge, the electron gun of the previous measurements was redesigned, using a negative beam-forming electrode as indicated in Fig. 6(c). The boundary conditions at the edge of the cathode are met by using a 122.5-degree swept-back edge and an electrode shaped and biased to bring a 67.5 degree cathode equipotential to the cathode edge.

Fig. 7 shows a series of photographs of the image obtained from a pin-hole camera using this gun. It is evident that the flow is much closer to the ideal. This tube, however, gave about 20 per cent greater than the design current at the normal operating potential (1,000 volts). At first it was believed that contributions from the conical sides of the cathode might be responsible for this, but several tubes made with attempts to mini-

mize emission from this region also gave high perveance. It seems probable that the extra current is caused by the reduction of the space charge density within the gun due to thermal-velocity spreading of the beam, and that this is a natural consequence of thermal velocities. In the previous gun this effect was more than compensated for by the extra beam convergence described. This explanation is enforced by the fact that the current does not follow the 3/2 power law of voltage but gives higher than predicted current at low voltages, approaching the design perveance at high voltages.

The arrangement indicated in Fig. 6(b) offers another solution. In this case the ideal boundary condition may not be met, but it is structurally more desirable than Fig. 6(c) and gave substantially the same performance, the thermal-velocity effects entirely masking any geometrical disturbance.

This design was used for gun No. 2 (Table I) and gave the beam cross section and transverse velocity distribution of Fig. 8. The transverse velocity distribution is Maxwellian, and increased in width over the distribution at the cathode. The magnitude of this increase corresponds to within 10 per cent of the theoretically predicted increase, provided the beam magnification is obtained from measured current density distribution.

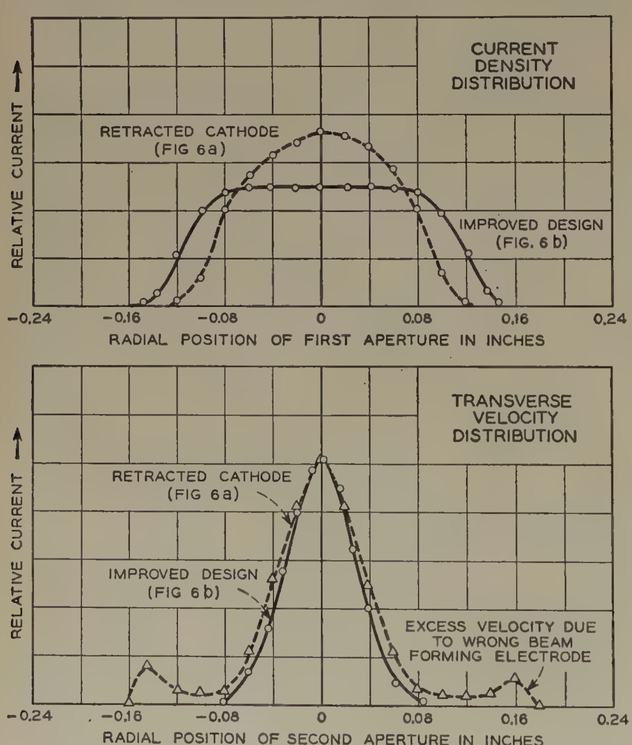

Fig. 8—Beam density and transverse-velocity distribution for gun No. 2, using the beam-forming electrode design of Fig. 6(b). The dashed curve is the result obtained with the design of Fig. 6(a).

It is to be expected that, besides determining the transverse velocity distribution, thermal effects would cause a widening of the beam and reduce the definition of the beam edge. A first order calculation of this is made in the companion paper,[1] giving for gun No. 2 the curve of Fig. 9. It is apparent that the beam diameter is significantly larger than the design value. This is believed to be due to the approximate nature of the relationship used in accounting for the divergence of the exit-aperture lens. When the actual mean beam diameter is used, the thermal-velocity spreading calculated is very close to the measured value.

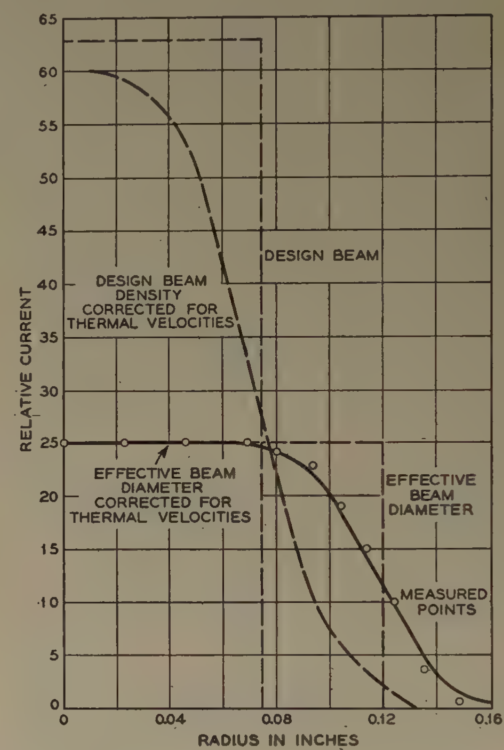

Fig. 9—Comparison of beam spread with calculated values for gun No. 2. The beam diameter is much greater than the design value, presumably due to the approximation involved in evaluating the exit lens. For the resulting beam size and density the thermal spreading is almost exactly the predicted value.

Other Gun Defects

In studying these guns several other defects were observed. The most common and obvious one is that of nonuniformity of cathode emission. This may be caused by poor cathode-coating adherence, granularity of the cathode surface, or deactivation of certain areas of the cathode by positive-ion bombardment. Tubes suffering from these troubles were photographed and the images are shown in Fig. 10 on the following page.

The non-emitting center area of Fig. 10(b) is usually seen as a tube approaches the end of its life and coincides with the development of a hole in the cathode through the coating and even into the base metal itself. The development of this spot very early in the life of the tube can be observed by cooling the cathode until the emission starts to be temperature-limited. To get a stable measurement of the effect of such a spot on focusing, the coating was deliberately removed from the center 7 per cent of the cathode area of gun No. 2, and the resulting beam measured. It is evident from Fig. 11 that an excess of inward velocity is developed, resulting in large cross over of electrons which must

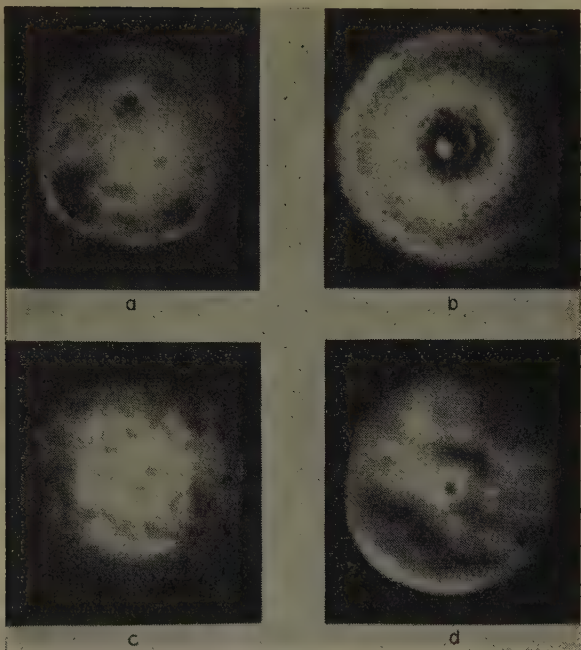

Fig. 10—Photographs of the images from several pin-hole camera tubes showing types of nonuniform emission. (a) Poor coating adherence. (b) Cathode with spot produced by back bombardment of positive ions. The ring is due to improper beam-forming electrode position. (c) Granularity of emitting surface (obtained by cooling the cathode of Fig. 7). (d) Image of cathode of Fig. 3 after 4,000 hours.

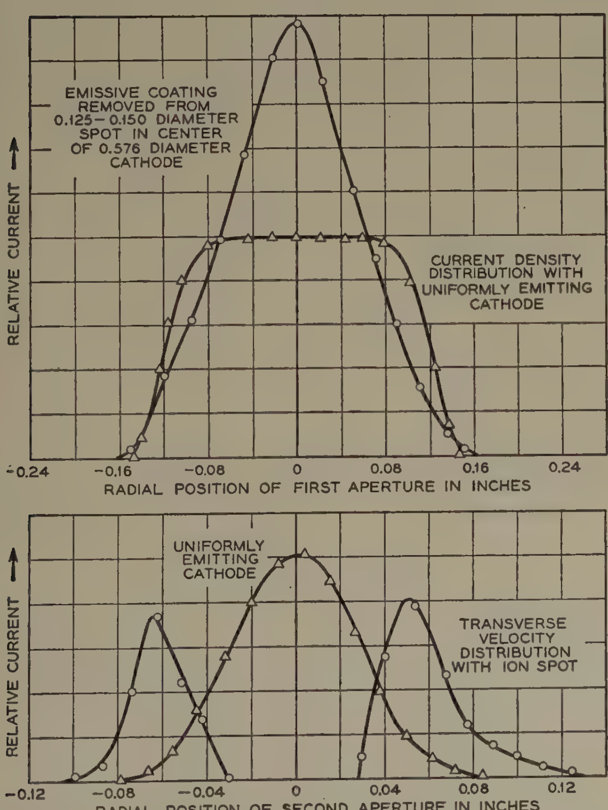

Fig. 11—Current density and velocity distribution of gun No. 2 with 7 per cent of emitting area at the cathode center removed to simulate an ion spot. Even though the beam appears to be more convergent in the gun, the anode interception was increased from 2 per cent to 14 per cent.

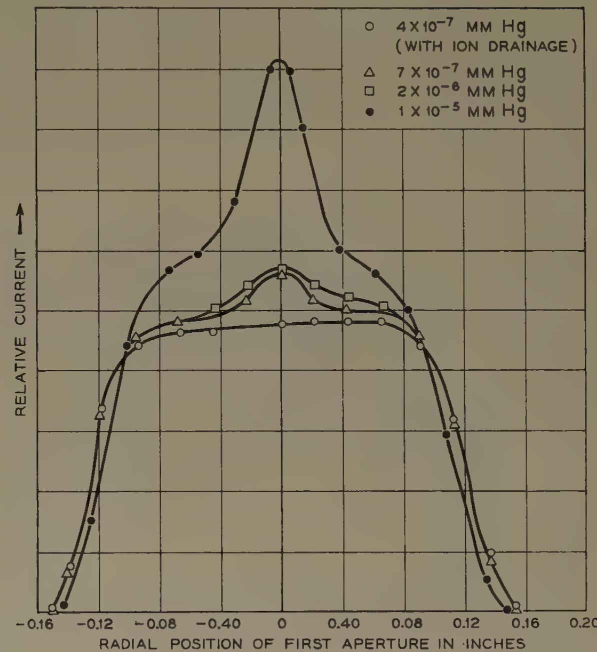

Fig. 12—Current-density distribution from gun No. 2 at various gas pressures, showing the effect of space-charge neutralization by positive ions.

cause a degradation of focusing. The most obvious effect near the anode aperture is one of a peaking of the current due to the greater convergence. Note, however, that the current outside of a certain diameter actually is increased. As a result of this, the anode interception actually increased from 2 per cent to 14 per cent with the introduction of the inactive area.

Effect of Pressure

The design of the Pierce electron gun also assumes a complete absence of positive ions in the beam. However, at pressures often considered good ($10^{-7}-10^{-8}$ mm Hg) sufficient ions are created to alter the space-charge conditions within the beam. To get a measure of the focusing properties of the gun in the absence of ionic effects, a special pulse technique was employed. The gun was pulsed with a 700-microsecond pulse at a 60-cycle repetition rate, and current measurements were made within the first 2–5 μsec of the pulse. Comparing these measurements with the currents at the end of the pulse, or with the dc observations, a clear measure of the effect of positive ions is obtained. Alternatively it was found that ion effects could be removed by applying a slight longitudinal voltage gradient beyond the anode of the gun to prevent the accumulation of ions in this region.

Fig. 12 shows the current density distribution obtained at various pressures and also the current density distribution obtained in the absence of appreciable ionic effects. When the pressure is increased, it is noted that the current density distribution becomes peaked in the center and that the beam diameter is very slightly

reduced. These curves show clearly that the positive ions fall into the potential depression in the center of the beam and partially neutralize the electronic space charge of the beam, resulting in a further convergence of the electrons.

Conclusions

The pin-hole electron camera is a powerful tool with which to study gun aberrations and emission phenomena. The electron-gun studies described show some of the common defects in electron focusing. In particular they emphasize the importance of extreme care in locating the beam-forming electrode to avoid serious aberrations of the electrons emitted near the edge of the cathode. They also show that positive-ion bombardment of the cathode in higher perveance guns has serious consequences in the electron flow long before it seriously degrades the gun perveance. The presence of positive ions in the stream causes extra beam convergence and consequently, a greater transverse velocity distribution.

The transverse thermal-velocity distribution is a more fundamental limitation to the flow. The measurements show that converging the electron beam widens the transverse thermal-velocity distribution in a predictable fashion, simply due to the geometry of flow. It is also indicated that especially in low-perveance low-voltage guns, the thermal velocities widen the beam and increase the perveance. Measurements indicate that when other defects are eliminated, the diverging effect of the ungridded anode aperture is significantly greater than predicted by the approximate formula.

Acknowledgment

Credit is due to several of our colleagues at the Bell Telephone Laboratories for helpful advice and stimulation in this work; and especially to L. J. Heilos, F. Krastin, H. T. Closson, W. A. Warne, and G. J. Stiles for their assistance.

Glossary of Symbols

i	Current density.
r	Radius measured in beam.
r_i	Radius measured in image.
v_c	Transverse velocity of electrons at any point at the cathode surface.
v_r	Radial velocity of electrons at r.
θ	Half-angle of convergence of electron guns.
ϕ	Angular direction.
e	Electron charge, 1.602×10^{-19} coulomb.
M	Beam convergence factor. M^2 is ratio of current density at cathode to that in beam.
m	Electron mass, 9.11×10^{-31} kilogram.
k	1.380×10^{-23} Joules/degree (Boltzman's constant).
T	Temperature, degrees Kelvin.
V_0	Voltage of anode, aperture, drift space and screen.
V_T	Voltage equivalent of transverse velocity.
I_0	Total current, amperes.
B	Magnetic field, gauss.
z	Distance measured from cathode, inches.
z_i	Separation between pinhole and image plane.
J_c	Current density in the z-direction at the cathode.
J_i	Current density in the image plane.

Thermal Velocity Effects in Electron Guns*

C. C. CUTLER†, FELLOW, IRE, AND M. E. HINES†, MEMBER, IRE

Summary—**A method is presented for determining the effect of the transverse velocities of electron emission on the flow in accelerated electron streams. Complete expressions are derived giving the degree and character of the beam spread in a Pierce-type electron gun resulting from the transverse velocity distribution, and an expression is derived for the image magnification in the pin-hole camera beam analyzer described in a companion paper.[1] The theory clearly shows the increase in effective (transverse) beam temperature caused by beam convergence.**

I. Introduction

Modern microwave tubes often require electron beams of small diameter and high current density. There are a number of factors which tend to limit the focusing in such beams, the most serious arising from the thermal-velocity distribution of the emitted electrons. It is common practice to use electron guns which converge the beam from a large cathode in order to obtain more current density than is available from practical emitters. This further accentuates the effect of the thermal velocities which ultimately limit the amount of convergence obtainable.

In this paper, a number of the effects of the thermal velocities of emission in a Pierce-type electron gun are treated theoretically. First, in Section II, a method is developed for determining the departure, caused by an initial transverse component of velocity of an electron from its ideal laminar path. This method is applied to the accelerating region, in an electron gun and in a field-free drift space under the influence of space charge. From this, in Section III, the beam spreading due to thermal velocities is determined, and the curves given make it easy to apply the results to any particular gun and beam dimensions. This concludes with the curves of Fig. 6 which show how the current density varies across the beam. The character of the variation is given in terms of the normal trajectory deviation σ which can be determined from the gun design parameters with the help of Fig. 5 in specific cases or through (32) in more general terms. Fig. 7 in turn gives the per cent current transmission within any radius, and Fig. 8 the current density contours for a specific case. A solution of uniform diameter focusing in a magnetic field for a beam dispersed by thermal velocities has been worked out by Pierce and Walker.[2] The solutions worked out in this paper, while similar in shape to the thermal distributions arrived at by these authors, do not constitute a solution to their boundary problem, nor is it clear that such a solution exists. The value of their work is in predicting a limiting minimum perturbation due to thermal velocity effects in magnetically focused electron flow.

This approach was initially worked out in an effort to calculate the image magnification of the pin-hole camera, shown in Fig. 1 which is useful in experimental study of electron gun aberrations.[1] The results are readily applicable to this problem as shown in Section IV. Eq. (36) gives the image magnification in terms of the tube dimensions and gun design parameters, with the help of Figs. 2 and 3.

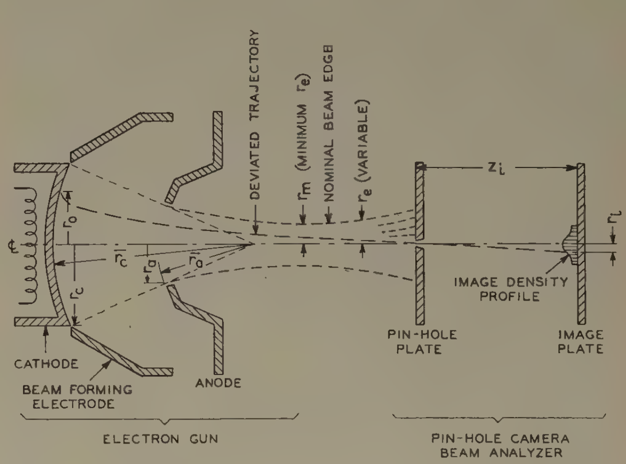

Fig. 1—Geometry of Pierce-type converging-electron gun and pin-hole camera beam analyzer.

A consequence of the convergence of an electron beam is an increase in the width of the transverse velocity distribution, the transverse velocities varying in inverse proportion with the beam diameter. A proof of this is inherent in the analysis, and it is shown in Section V that the apparent increase in velocity spread is simply a consequence of the geometry of the electron flow.

Partial verification of the conclusion is indicated by measurements presented in a companion paper.[1] The check appears to be reasonably close provided that an additional correction is made to account for the excess divergence resulting from the rather large hole in the anode apertures[3] of the guns tested. Throughout the paper, MKS units will be used.

II. Electron Trajectories

We shall be almost exclusively concerned with the converging type of Pierce electron gun illustrated in Fig. 1, though the development in (1) through (8) is

* Original manuscript received by the IRE, October 18, 1954; revised manuscript received December 2, 1954.

† Bell Telephone Laboratories, Murray Hill, N. J.

[1] C. C. Cutler and J. A. Saloom, "Pin-hole camera investigation of electron beams," Proc. I.R.E., vol. 43, pp. 299–306; March, 1955.

[2] J. R. Pierce and L. R. Walker, "Brillouin Flow with Thermal Velocities," *Jour. Appl. Phys.*, vol. 24, pp. 1328–1330; October, 1953.

[3] Pierce's design theory accounts for the anode aperture lens effect by use of the standard small-aperture formula for the focal length. This formula becomes somewhat inaccurate when the diameter of the aperture is an appreciable fraction of the cathode-anode spacing.

more general. The theory of this gun has been presented by Pierce[4] for the limiting case of negligible initial electron velocities and small angles of convergence. The beam starts with space-charge limited flow from a cathode which is a concave section of a sphere. The anode electrode at high potential and the beam-forming electrode at cathode potential are so shaped that in the accelerating region the electric-field pattern within the beam matches as closely as possible the fields which would exist between complete concentric spherical electrodes with space-charge limited flow through the interelectrode space.

In a practical gun, the anode is apertured for passage of the beam. The field pattern in the immediate vicinity of the aperture is assumed to act as an ideal thin diverging lens which causes the beam to be less convergent after passing through the aperture. Beyond the anode, it is assumed that the only fields acting are those resulting from the space charge of the beam itself. The space charge fields produce a radial repulsive force which prevents the beam from achieving a point focus. Instead, the beam converges to a minimum diameter beyond the anode and then diverges again. It is common practice in electron tubes to inject the beam into a uniform parallel magnetic field at a point near its minimum diameter in an attempt to focus the beam in the uniform-diameter "Brillouin Flow" condition.[5]

In the gun design theory of Pierce, the electron flow is assumed to be essentially laminar in character with no crossing of paths. We assume this for all electrons except the one whose trajectory across normal paths we wish to determine. The paraxial-ray assumptions made in dealing with the flow are: (1) all angles of convergence and divergence of laminar flow lines are small, (2) longitudinal electric fields are uniform over a given cross section of the beam, (3) all radial fields are proportional to the radius from the axis. As a consequence of these assumptions, radial velocities of the laminar flow electrons are proportional to the radius from the axis, and the axial velocities are uniform over a given cross section.

It is useful to employ a radial coordinate μ which varies from zero on the beam axis to unity at the beam edge. Electrons with zero initial velocity follow paths of constant μ. Let r_e be the variable radius of the edge of the beam ($\mu=1$) and r be the actual radial position of a particular electron crossing the surfaces of constant μ, where r as a function of the axial coordinate z would describe the path of the wandering electron. We write

$$r = \mu r_e. \tag{1}$$

Differentiating twice gives

$$\frac{d^2r}{dt^2} = r_e\frac{d^2\mu}{dt^2} + 2\frac{d\mu}{dt}\frac{dr_e}{dt} + \mu\frac{d^2r_e}{dt^2}. \tag{2}$$

[4] J. R. Pierce, "Theory and Design of Electron Beams," 2nd. ed., D. Van Nostrand Company, pp. 173–193; 1954.

[5] *Ibid.*, pp. 152–157.

By the paraxial ray assumption (item 3 above),

$$\frac{d^2r}{dt^2} = \mu\frac{d^2r_e}{dt^2}; \tag{3}$$

therefore,

$$r_e\frac{d^2\mu}{dt^2} = -2\frac{dr_e}{dt}\frac{d\mu}{dt}. \tag{4}$$

Rearranging terms results in

$$\frac{d\left(\dfrac{d\mu}{dt}\right)}{\dfrac{d\mu}{dt}} = -2\frac{dr_e}{r_e}. \tag{5}$$

A single integration gives

$$\ln\frac{d\mu}{dt} = -\ln r_e^2 + \ln r_{e_1}^2 + \ln\left(\frac{d\mu}{dt}\right)_1, \tag{6}$$

where the subscript 1 indicates the value at any given cross-section plane. In general we will later refer to the cathode and anode plane s by c and a, respectively. Taking antilogarithms gives

$$\frac{d\mu}{dt} = \left(\frac{d\mu}{dt}\right)_1\frac{r_{e_1}^2}{r_e^2}. \tag{7}$$

A second integration gives

$$\mu - \mu_1 = \left(\frac{d\mu}{dt}\right)_1\int_{t_1}^{t}\frac{dt}{\left(\dfrac{r_e}{r_{e_1}}\right)^2}. \tag{8}$$

This is the fundamental equation for thermal trajectory tracing. The integral on the right involves terms which are presumably known for the ideal laminar trajectories. It is apparent that if space-charge forces are appreciable, the actual trajectories must approximate the ideal in order to justify the above derivation. It should be noted that all the above equations apply for any geometry for which the listed assumptions are valid.

Trajectories in the Accelerating Region of a Pierce Gun

The integral of (8) is to be taken from the cathode through the accelerating region of the gun and beyond to any point in the beam-spread region desired. That portion of the integral for the region of acceleration is taken up in this section. For this region the limits of the integration are t_c and t_a, respectively, c referring to values at the cathode, and a to values at the anode aperture.

Langmuir's expression for the space charge limited current in a complete spherical diode is

$$I = \frac{16\pi\epsilon\sqrt{2\eta}\,V^{3/2}}{9(-\alpha)^2}; \tag{9}$$

where $(-\alpha)^2$ is a tabulated function[6] of the ratio of the

[6] *Ibid.*, p. 183.

spherical cathode to anode radii, and $\eta=e/m$ of the electron. V is the potential difference between the cathode and anode, and ϵ is the dielectric constant of free space (see Glossary of Symbols). For a conical section of a sphere as used in a Pierce gun of half-angle θ, the beam current I_0 is given by

$$I_0 = \frac{16\pi\epsilon\sqrt{2\eta}\, V^{3/2} \sin^2\left(\frac{\theta}{2}\right)}{9(-\alpha)^2}. \tag{10}$$

Eqs. (9) and (10) apply for any ratio of radii. If one assumes the current to be constant, these equations may be used to define the variation of potential with distance from the cathode surface in terms of the function $(-\alpha)^2$. Thus,

$$V = V_a \frac{(-\alpha)^{4/3}}{(-\alpha_a)^{4/3}} \tag{11}$$

where a indicates values at the anode. In an increment of time dt, an electron will move a distance $d\bar{r}$ at a velocity $\sqrt{2\eta V}$, or

$$dt = -\frac{d\bar{r}}{\sqrt{2\eta V}} = -\frac{\bar{r}_c(-\alpha_a)^{2/3} d\left(\frac{\bar{r}}{\bar{r}_c}\right)}{\sqrt{2\eta V_a}\,(-\alpha)^{2/3}}, \tag{12}$$

where $\bar{r}_c$ is the radius of curvature of the cathode and $\bar{r}$ is the variable distance from the same center. Substituting into the integral of (8) and referring to the cathode ($r_{e1}=r_c$), we obtain

$$\int_{t_e}^{t_a} \frac{dt}{\left(\frac{r_e}{r_c}\right)^2} = -\frac{\bar{r}_c(-\alpha_a)^{2/3}}{\sqrt{2\eta V_a}} \int_1^{\bar{r}_a/\bar{r}_c} \frac{d\left(\frac{\bar{r}}{\bar{r}_c}\right)}{\left(\frac{r_e}{r_c}\right)^2 (-\alpha)^{2/3}}. \tag{13}$$

Substituting $\bar{r}/\bar{r}_c$ for r_e/r_c (which follows from radial flow) and inverting the variable of integration simplifies the expression, yielding

$$\int_{t_c}^{t_a} \frac{dt}{\left(\frac{r_e}{r_c}\right)^2} = \frac{\bar{r}_c(-\alpha_a)^{2/3}}{\sqrt{2\eta V_a}} \int_1^{\bar{r}_c/\bar{r}_a} \frac{d\left(\frac{\bar{r}_c}{\bar{r}}\right)}{(-\alpha)^{2/3}}. \tag{14}$$

Fig. 2 shows a plot of the right-hand integral as a function of the upper limit. From (8) and (14) we can determine the trajectory within the accelerating region of the gun for an electron with a given initial transverse velocity. We shall apply this to the beam spread problem in Section III.

Trajectories in a Drift Region

In the drift region beyond the anode aperture, the beam diameter varies under the influence of the diverging forces of the space charge within the beam. For the case of a shielded beam of slowly varying diameter, we may approximate by using the equations applicable for cylindrical geometry,[7]

$$\frac{1}{\eta}\frac{d^2 r_e}{dt^2} = \text{Radial field strength} = \frac{\rho}{2\pi\epsilon r_e},$$

where ρ is the linear charge density in coulombs per meter of length. In terms of the beam current and voltage this becomes

$$\frac{d^2 r_e}{dt^2} = \frac{\eta I_b}{2\pi\epsilon r_e\sqrt{2\eta V_a}}. \tag{15}$$

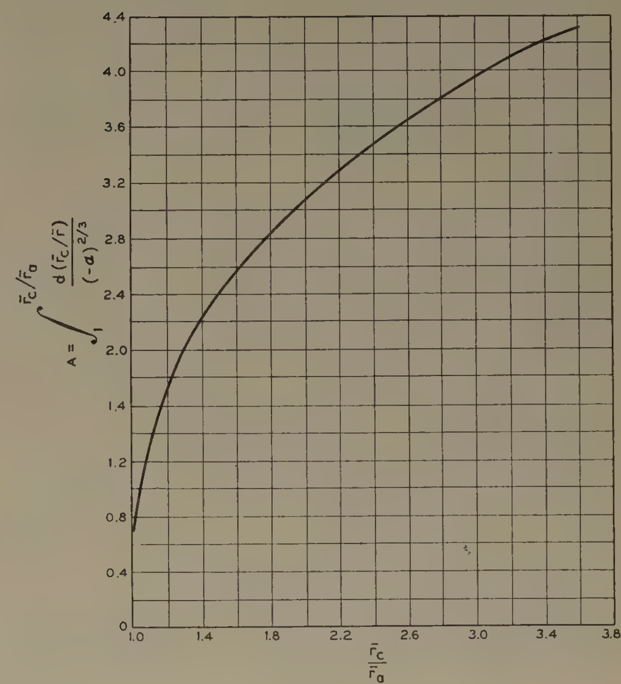

Fig. 2—A curve useful in calculating electron trajectories in the accelerating region of Pierce-type electron guns.

Substituting for I_b from (10), and assuming θ to be small so that we may approximate the sine of an angle by the angle we obtain,

$$\frac{d^2 r_e}{dt^2} = \frac{2\eta V_a r_c^2}{9(-\alpha_a)^2 \bar{r}_c^2}\frac{1}{r_e}. \tag{16}$$

This equation may be put into integrable form by substituting

$$\omega = \frac{dr_e}{dt}, \qquad \omega\frac{d\omega}{dr_e} = \frac{d^2 r_e}{dt^2}.$$

Integrating in this manner gives:

$$\omega = \frac{dr_e}{dt} = \pm\frac{\sqrt{2}\sqrt{2\eta V_a}}{3(-\alpha_a)}\frac{r_c}{\bar{r}_c}\left(\ln\frac{r_e}{r_m}\right)^{1/2}, \tag{17}$$

which includes the boundary condition that dr_e/dt is

[7] *Ibid.*, p. 147.

zero at the minimum beam radius r_m. The plus sign applies for a diverging beam and the minus sign for a converging beam. The second part of the integral for (8) covering the drift region, is therefore

$$\int_{t_a}^{t} \frac{dt}{\left(\frac{r_e}{r_c}\right)^2} = \bar{r}_c \frac{r_c}{r_m}\left(\frac{3(-\alpha_a)}{\sqrt{2}\sqrt{2\eta V_a}}\right) \cdot \int_{r_a/r_m}^{r_e/r_m} \frac{\pm d\left(\frac{r_e}{r_m}\right)}{\left(\frac{r_e}{r_m}\right)^2\left(\ln\frac{r_e}{r_m}\right)^{1/2}} \cdot \qquad (18)$$

We may change the variable of integration to u, defined by

$$\frac{r_e}{r_m} = e^u, \qquad (19)$$

which yields

$$\int_{t_a}^{t} \frac{dt}{\left(\frac{r_e}{r_c}\right)^2} = 3\bar{r}_c \frac{r_c}{r_m} \frac{\sqrt{\frac{\pi}{2}(-\alpha_a)^2}}{\sqrt{2\eta V_a}} \cdot \left[\frac{2}{\sqrt{\pi}}\int_0^{\sqrt{\ln r_a/r_m}} e^{-u^2}du \pm \frac{2}{\sqrt{\pi}}\int_0^{\sqrt{\ln r_e/r_m}} e^{-u^2}du\right]$$

$$= 3\bar{r}_c \frac{r_c}{r_m} \frac{\sqrt{\frac{\pi}{2}(-\alpha_a)^2}}{\sqrt{2\eta V_a}} \cdot \left\{\operatorname{erf}\sqrt{\ln (r_a/r_m)} \pm \operatorname{erf}\sqrt{\ln (r_e/r_m)}\right\}. \qquad (20)$$

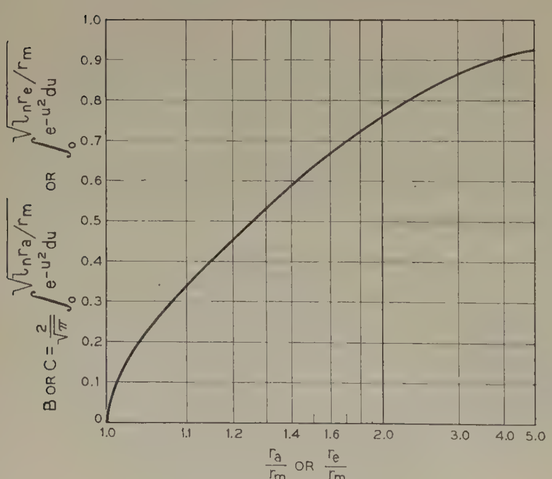

Fig. 3—A curve useful in calculating beam divergence in a drift region.

This completes the integral of (8). The plus sign applies for points beyond the beam minimum and the minus sign for points before the beam minimum is reached. Values of the error functions are given for convenience in Fig. 3. The value of r_e as a function of distance can be obtained from the gun design data and the universal beam spread curve.[8]

Complete Trajectory, Gun Plus Drift Region

We are now in a position to write the complete expression for the trajectory deviation of an electron which has a known initial transverse velocity and which remains within the beam. From (1) and (8) it is apparent that an electron will deviate from its nominal position by an amount Δr, given by

$$\Delta r = \left(\frac{dr}{dt}\right)_c \frac{r_e}{r_c} \int_{t_c}^{t} \frac{dt}{\left(\frac{r_e}{r_c}\right)^2} \cdot \qquad (21)$$

Assembling the integral from (14) and (20), we have

$$\frac{\Delta r}{\left(\frac{dr}{dt}\right)_c} = \frac{\bar{r}_c}{\sqrt{2\eta V_a}}\left[\frac{r_e}{r_c}(-\alpha_a)^{2/3}\int_1^{\bar{r}_c/\bar{r}_a} \frac{d\left(\frac{\bar{r}_c}{\bar{r}}\right)}{(-\alpha)^{2/3}} + 3\frac{r_e}{r_m}\sqrt{\frac{\pi}{2}(-\alpha_a)^2}\left(\operatorname{erf}\sqrt{\ln\frac{r_a}{r_m}} \pm \operatorname{erf}\sqrt{\ln\frac{r_e}{r_m}}\right)\right]. \qquad (22)$$

An electron starting with an initial transverse velocity $(dr/dt)_c$ at the cathode, will arrive at a later point, removed from the line of laminar flow by a distance Δr given by (22). The positive sign in this expression should be taken if the cross section of interest is beyond the minimum beam diameter of the space-charge beam-spread curve, and the negative sign in case the plane of interest comes before the plane of the minimum beam diameter.

This equation is complete for a single electron. The integral and the erf functions are evaluated in Figs. 2 and 3. We apply it to the case of many electrons under the influence of thermal velocities in the next section, and discuss its evaluation further at the end of that section.

III. Thermal Velocity Beam Spreading

The theory of the preceding section applies to a single wandering electron which remains within an otherwise ideally focused beam. Though the theory was applied to the case of an initial transverse velocity in the radial direction only, it may also be shown to apply for off-axis electrons with initial velocities in any transverse direction. This is a consequence of the linear character of the variation of the radial fields with radius under the paraxial-ray assumption. For axially symmetric systems in which $E_r = ar$ it is fully equivalent to write $E_x = ax$ and $E_y = ay$ in a Cartesian co-ordinate. Thus in (1) through (8) we could replace r by x or y and μ by

[8] *Ibid.*, p. 150.

μ_x or μ_y with full validity. This may be carried through to (21). Thus the magnitude of our deflection is independent of the direction of the initial velocity. The theory is only approximate for an electron which leaves the nominal beam boundary. To obtain the effect on the over-all beam we assume that each electron moves independently as calculated in the field of an ideal beam, and then we sum the effect of all electrons. If the beam is appreciably spread by these effects or becomes distorted by other aberrations, considerable error may result. Nevertheless, the theory does provide a first-order calculation of beam spreading due to thermal velocities which should be reasonably accurate unless the effect is very large. It should be noted that usually we are most concerned with the distribution of electrons outside of the nominal beam boundary and that the inaccuracies of our theory for these electrons are in the direction of predicting greater trajectory deviations than should actually occur in practice.

A Maxwellian distribution of the transverse velocities is assumed for the emitted electrons. The transverse components of the velocity distribution are specified by the relation

$$dJ_c = J_c \frac{m}{2\pi kT} e^{-(mv_x^2/2kT)-(mv_y^2/2kT)} dv_x dv_y; \tag{23}$$

where J_c is the current density in the z-direction at the cathode surface, T is the cathode temperature in degrees Kelvin, m is the mass of the electrons, and k is Boltzmann's constant.

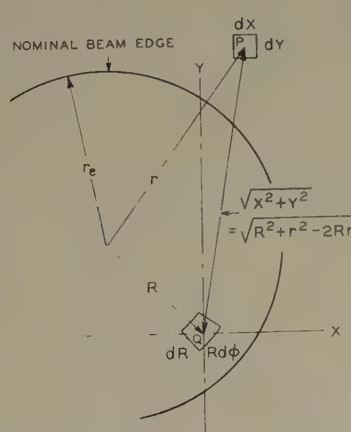

Fig. 4—Geometry for calculation of the beam spread caused by thermal velocities.

Fig. 4 shows the geometry we are considering at a cross-section of the beam an arbitrary distance from the cathode. A current element nominally expected to arrive at Q will be distributed in a symmetrical zone around Q and some of this current will arrive at the point of interest P. This point is at a radius r from the center of the beam and the beam would have a radius r_e in the absence of thermal velocities.

It is convenient at this point to define a normal deviation σ where

$$\sigma = \frac{\Delta r}{\left(\dfrac{dr}{dt}\right)_c} \sqrt{\frac{kT}{m}}, \tag{24}$$

in which Δr is the trajectory deviation of an electron with an initial transverse velocity $(dr/dt)_c$. It follows that

$$\sigma = \frac{\Delta r}{\left(\dfrac{dr}{dt}\right)_c} \sqrt{\frac{kT}{m}} = \frac{x}{v_x}\sqrt{\frac{kT}{m}} = \frac{y}{v_y}\sqrt{\frac{kT}{m}}, \tag{25}$$

where x and y are distances of deviation as shown in Fig. 4 and $\Delta r/(dr/dt)_c$ can be evaluated from (22). [See (32).]

Referring to Fig. 4, that part of the current dI_p which in the absence of thermal velocities would fall in the area $dx\,dy$ at P is given by substituting x, y, and σ into (23).

$$dI_p = \frac{J_0 R\,dR\,d\Phi}{2\pi} e^{-(x^2+y^2)/2\sigma^2} d\left(\frac{x}{\sigma}\right) d\left(\frac{y}{\sigma}\right), \tag{26}$$

where J_0 is the current density in the beam in the absence of thermal effects. That part of the current density at P from the area at Q is given by

$$dJ_p = \frac{dI_p}{dx\,dy} = \frac{J_0}{2\pi\sigma^2} e^{-(x^2+y^2)/2\sigma^2} R\,dR\,d\Phi. \tag{27}$$

To integrate this expression, we eliminate x and y by the law of cosines,[9]

$$x^2 + y^2 = R^2 + r^2 - 2Rr\cos\Phi. \tag{28}$$

The total current density at P is determined by integrating over the area of the nominal beam circle. That is,

$$J_P = \frac{J_0}{2\pi\sigma^2} e^{-r^2/2\sigma^2} \int_0^{r_e} R e^{-R^2/2\sigma^2} dR \int_0^{2\pi} e^{+(Rr\cos\Phi)/\sigma^2} d\Phi. \tag{29}$$

The second integral in (29) may be identified as a representation of the modified Bessel function

$$\int_0^{2\pi} e^{(Rr\cos\Phi)/\sigma^2} d\Phi = 2\pi I_0\left(\frac{rR}{\sigma^2}\right). \tag{30}$$

Thus we get, for the ratio of the expected current density J at radius r to the idealized current density J_0,

$$\frac{J_r}{J_0} = e^{-r^2/2\sigma^2} \int_0^{r_e/\sigma} \left(\frac{R}{\sigma}\right) e^{-R^2/2\sigma^2} I_0\left(\frac{rR}{\sigma^2}\right) d\left(\frac{R}{\sigma}\right). \tag{31}$$

This is the relationship we have been working for. It gives the ratio of the beam-current density to the idealized value in terms of the beam radius and the position in the beam, and of parameter σ which can be obtained from the gun-design parameters. From (22) and (24)

[9] This method of integration was suggested by W. E. Danielson of the Bell Telephone Laboratories.

$$\sigma = \sqrt{\frac{kT}{m}}\frac{\bar{r}_c}{\sqrt{2\eta V_a}}\left[\frac{r_e}{r_c}(-\alpha_a)^{2/3}A + 3\frac{r_e}{r_m}\sqrt{\frac{\pi}{2}(-\alpha_a)^2(B \pm C)}\right], \quad (32)$$

or

$$\frac{r_e}{\sigma} = \frac{1}{\sqrt{\frac{kT}{2eV_a}}}\,\frac{\sin\theta}{(-\alpha_a)^{2/3}A + 3\frac{r_c}{r_m}\sqrt{\frac{\pi}{2}(-\alpha_a)^2(B \pm C)}}; \quad (33)$$

where A, B, and C are defined and plotted in Figs. 2 and 3. In (32) the minus sign is to be taken for points between the anode and the point of minimum beam diameter, and the plus sign is to be taken for points beyond.

In the Pierce gun-design theory, the degree of beam compression (r_m/r_c) depends only upon the ratio of the cathode and anode radii of curvature ($\bar{r}_c/\bar{r}_a$).[10] For the position at the nominal minimum beam diameter ($r_e=r_m$), and also at the anode aperture ($r_e=r_a$), the bracketed expression in (32) is a function of the ratio $\bar{r}_c/\bar{r}_a$ only, and the equation can be solved to give the simple relation plotted in Fig. 5. It should be noted that for guns with fairly large angles (θ) of convergence, or wherein the anode aperture compares in size with the cathode-anode spacing, values of r_m/r_c may differ somewhat from that given by the design curve of Pierce because of second-order lens effects at the aperture. In such cases, (32) still may be calculated provided that the correct strength of the exit lens can be determined, and from this the value and position of r_m.[11]

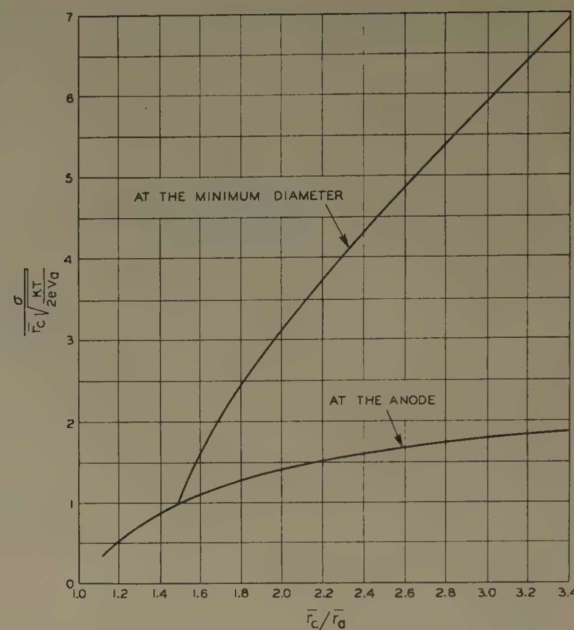

Fig. 5—Curves showing the transverse displacement of electron trajectories in Pierce-type guns. These curves may be interpreted as giving the normal displacement σ for a known value of cathode temperature.

Eq. (31) has been evaluated by numerical integration, and gives a family of functions (Fig. 6) of normalized radius r/r_e for various values of r_e/σ. In turn,

[10] *Ibid.*, p. 188, Figs. 10 and 12.

[11] It is pointed out in the companion paper that this effect was large in some of the guns tested.

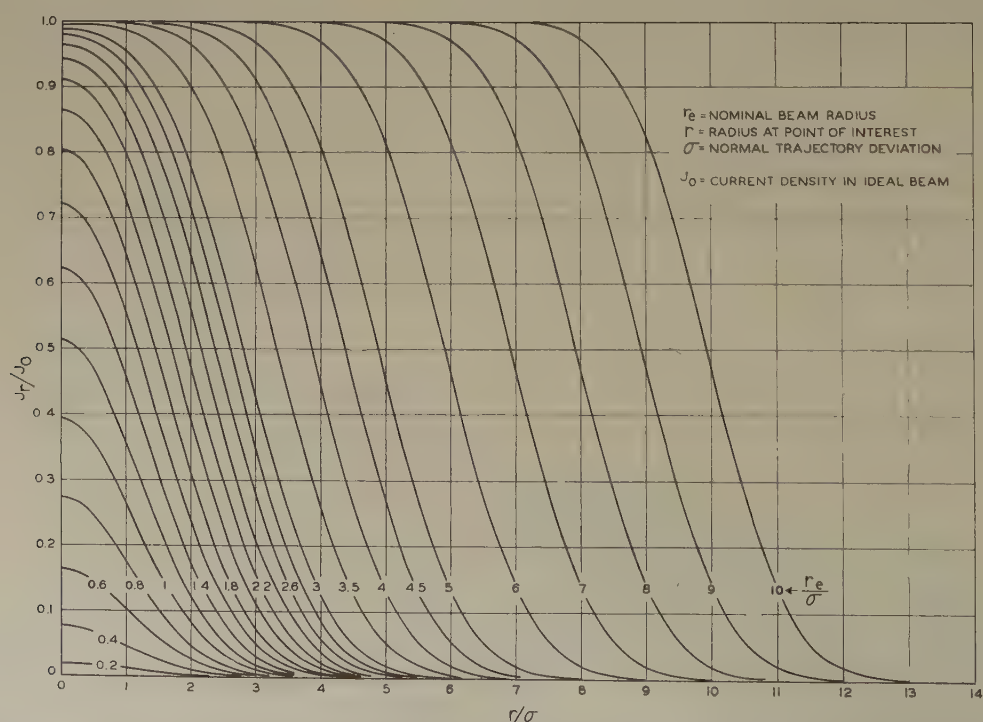

Fig. 6—Curves showing the character of the current density variation with radius in a beam which has been dispersed by thermal velocities. Here r_e is the nominal beam radius, r is the radius variable, and σ is the standard deviation defined in (38) and given in Fig. 4 for two specific positions.

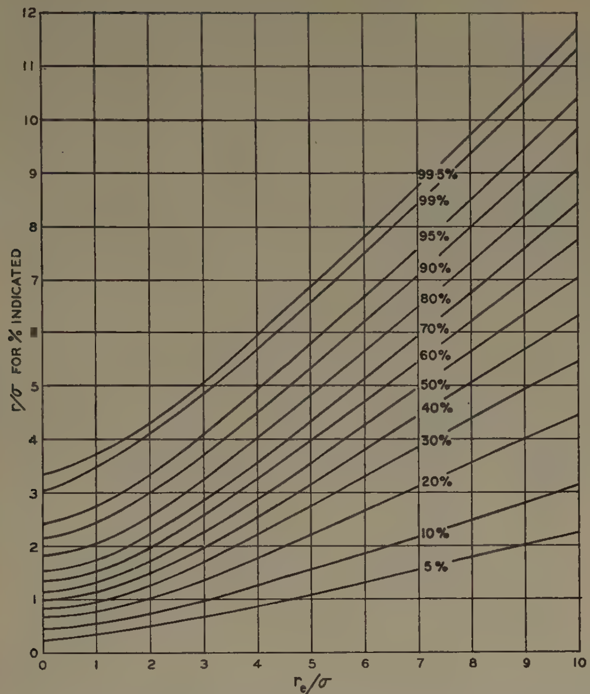

Fig. 7—Curves showing the per cent of the total beam current to be found within any given radius in a beam dispersed by thermal velocities as in Fig. 6.

Fig. 7 shows the normalized radius r/σ as a function of r_e/σ for various percentages of the total enclosed.

Fig. 8 shows how this theory predicts the variation in the current density profile of a beam with distance from the anode for a particular case. Profiles of current density vs radius are shown at a number of different points along the beam. The curve marked "Idealized Radius" is that predicted by Pierce's gun theory. The other curve shows the radius enclosing 99 per cent of the current as predicted by this theory. An important consequence to be noted is that the diameter enclosing 99 per cent of the current reaches a minimum at a point somewhat closer to the anode than the normal beam spread curve predicts.

IV. Image Magnification of the Pin-Hole Camera Tube

In the "pin-hole camera" beam analyzer described by Cutler and Saloom[1] the geometry is as shown in Fig. 1. An apertured plate with a single pin hole intercepts all but a very tiny part of the beam. The electrons, which pass, spread out according to the transverse velocity distribution and produce a kind of image on the back plate. Theoretically, electrons arriving at any point in the image must have come from a particular corresponding point on the cathode, and must have started at just the right transverse velocity to bring them to the aperture. Under our idealized conditions, we would expect a complete image of the cathode but not a uniformly illuminated one. Instead, we would expect a Gaussian distribution of image intensity with a maximum corresponding to the point on the cathode from which electrons with zero initial velocity would pass the aperture. In some cases, noted experimentally by Cutler and Saloom, the Gaussian distribution falls off so rapidly that the edge of the cathode is not detectable. This is to be expected in fairly wide-angle guns of low compression. In other cases, they noted marked distortions of the expected Gaussian distribution which were caused by improper geometry or spotty cathode emission. Interpretation of such a distorted image requires a knowledge of the cathode-to-image magnification of the pinhole camera.

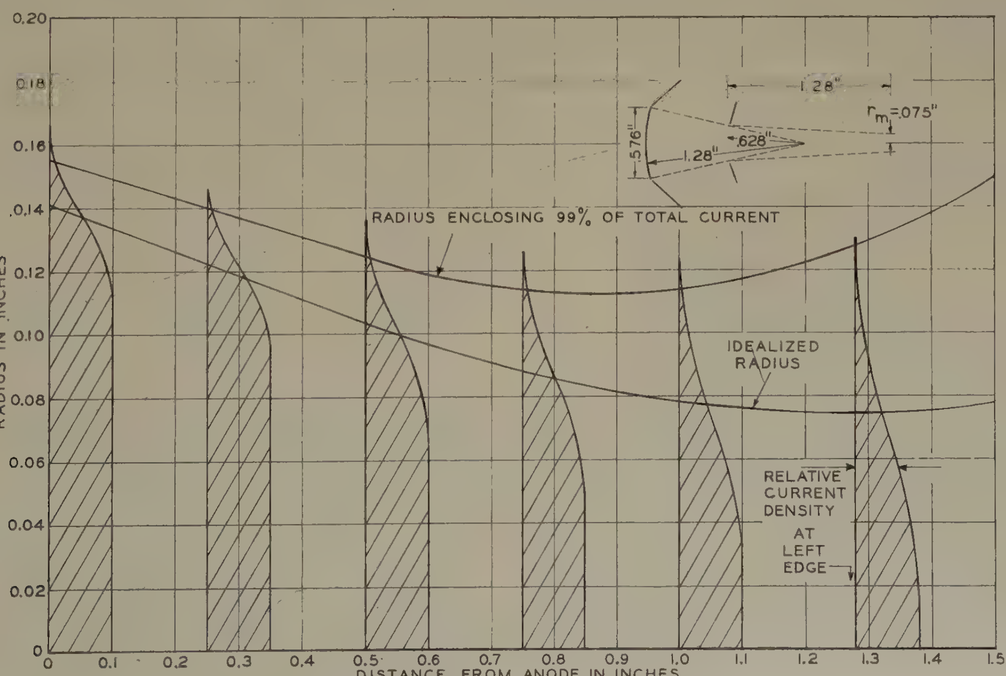

Fig. 8—Curves showing how the current density profile across the beam varies with distance from the anode for the particular gun geometry shown at an anode potential of 1,000 volts and a cathode temperature of 830 degrees C.

Referring to Fig. 1, it is apparent that the image height r_1 is given by

$$r_i = Z_i \frac{\left(\frac{dr}{dt}\right)_{ph}}{\left(\frac{dz}{dt}\right)_{ph}} = Z_i \frac{\left(r_e \frac{d\mu}{dt}\right)_{ph}}{\sqrt{2\eta V_a}}, \tag{34}$$

where ph refers to values at the pin-hole plane. We can write for the object height r_0,

$$r_0 = (\mu_{ph} - \mu_c) r_c,$$

which from (8) gives

$$r_0 = r_c \left(\frac{d\mu}{dt}\right)_c \int_{t_c}^{t_{ph}} \frac{dt}{\left(\frac{r_e}{r_c}\right)^2}. \tag{35}$$

This is the integral that we evaluated in Section III. Its value from t_c to t_a is given by (14) and Fig. 2. The value in the drift space, from t_a to t_{ph} is given by (20) and Fig. 3. Combining (34) and (35), and using (7) and the above mentioned integrals, gives for the image magnification:

$$\frac{r_i}{r_0} = \frac{\frac{Z_i}{\bar{r}_c}}{\frac{r_e}{r_c}\left[(-\alpha_a)^{2/3} A + 3\sqrt{\frac{\pi}{2}(-\alpha_a)^2}\,\frac{r_c}{r_m}\,(B \pm C)\right]}, \tag{36}$$

wherein r_e is the beam radius at the pin-hole plane. The denominator in (36) is identical to the bracketed expression in (32) or (22). If the pin-hole plane is at the anode aperture position or the nominal position of minimum beam diameter, it is the quantity plotted in Fig. 5. Otherwise using the tube and beam dimensions with (36) and Figs. 2 and 3 one can compute the image magnification.

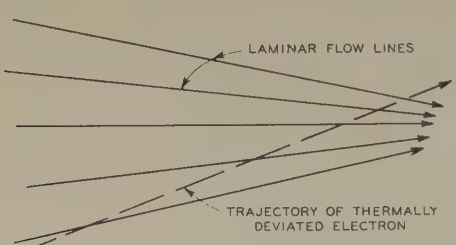

Fig. 9—Sketch illustrating how the transverse velocity of an electron across laminar flow lines is increased through beam convergence.

V. Beam Convergence and Beam Temperature[12]

In Fig. 9 we picture a converging electron stream in a unipotential region where all paths are straight. One electron is shown initially crossing the lines of the simple laminar flow at a slight angle, presumably because of an initial thermal velocity. It is apparent that this electron finds itself crossing laminar lines with an ever increasing angle as the beam converges. Hence it is evident that beam convergence causes an increase in the transverse velocity distribution. We may put this on a quantitative basis for the general case of accelerated streams through the use of (7).

$$\frac{d\mu}{dt} = \left(\frac{d\mu}{dt}\right)_c \frac{r_c^2}{r_e^2}.$$

Multiplying by r_e gives the transverse velocity of the wandering electron with respect to the coincident laminar path. From (7), then, we can write

$$\frac{r_e\left(\frac{d\mu}{dt}\right)}{r_c\left(\frac{d\mu}{dt}\right)_c} = \frac{v_{Te}}{v_{Tc}} = \frac{r_c}{r_e} = \frac{1}{M}. \tag{37}$$

The beam compression ratio M is the same as used by Pierce.[12] We have carried the argument for radial velocities only, but it applies equally well for either component of transverse velocity. It follows that if at the cathode the current density is distributed in transverse velocity as given by (24),

$$dJ_c = (\text{const}) e^{-m/2kT(v_x^2 + v_y^2)} dv_x dv_y. \tag{24}$$

At the plane z, the transverse velocities of electrons crossing any lamina are increased over those crossing any lamina at the cathode by the factor $1/M$. The new distribution of velocities is then

$$dJ_z = \text{const } e^{-(m/2kT)M_z^2(v_x^2 + v_y^2)} dv_x dv_y, \tag{38}$$

and to all intents the temperature has been effectively increased by $1/M_z^2$ where M_z is the ratio of the beam radius at z to that at the cathode. In the case of laminar flow M^2 is inversely proportional to the current density, which sometimes can be more specifically determined.

Gun aberrations and nonuniformities of emission including the limited size of the cathode, obviously affect the velocity distribution. These effects are superimposed upon the thermal effect which remains a basic limitation to the electron flow.

VI. Conclusion

The theory presented here points out the basic limitations in electron gun focusing which result from the thermal velocities of electron emission. The effects are most pronounced for highly convergent guns at low voltages where significant beam spreading and a marked increase in the thermal transverse velocities in the beam are to be expected. The equations and curves shown make it possible to predict the performance of practical guns, as has been shown by the experimental work described in this issue of the Proceedings by Saloom and Cutler.[1]

[12] *Ibid.*, p. 122. The effect of beam convergence on transverse velocities is indicated by J. R. Pierce from other considerations.

Glossary of Symbols

A The value of

$$\int_1^{\bar{r}_c/\bar{r}_a} \frac{d\left(\frac{\bar{r}_c}{\bar{r}}\right)}{(-\alpha)^{2/3}}.$$

a The value at the accelerating anode.

$(-\alpha^2)$ Langmuir's parameter for the solution of the spherical diode. (See reference 6.)

B The value of erf $\sqrt{\ln (r_a/r_m)}$.

C The value of erf $\sqrt{\ln (r_e/r_m)}$.

c The value at the cathode plane.

e Electronic charge $=1.602\times10^{-19}$ coulombs.

ϵ Dielectric constant of free space $=10^{-9}/36\pi$ farads per meter.

I_0 Total average current flow, amperes.

J Current density in the z direction.

J_0 Current density in the beam at position z in the absence of thermal effects.

k Boltzman's constant $=1.380\times10^{-23}$ Joules per degree.

m Electronic mass $=9.11\times10^{-31}$ kilograms.

η $e/m=1.759\times10^{11}$ coulombs/kilogram.

P A point to which electrons may flow under the influence of transverse velocities.

Q A point in the beam.

R Radial distance to Q.

r Radial position of an electron.

r_m Beam radius at the minimum of the beam-spread curve.

r_e Radius of the beam edge.

r_c Radius of cathode edge measured from axis.

$\bar{r}$ Radial dimension in spherical co-ordinates measured from the point of convergence of flow in a spherical diode.

$\bar{r}_c$ Radius of curvature of the cathode.

r_0 Object height in the pin-hole camera tube.

r_i Image height in the pin-hole camera tube.

Δr Trajectory deviation of an electron with an initial transverse velocity $(dr/dt)_c$.

ρ Linear charge density in coulombs per meter of length.

σ A normal deviation, defined as

$$\frac{\Delta r}{\left(\frac{dr}{dt}\right)_c}\sqrt{\frac{kT}{m}}.$$

T Cathode temperature, degrees Kelvin.

t Time variable.

θ Half angle of convergence in Pierce-type gun.

u A convenient variable $e^{u^2}=r_e/r_0$.

μ Ratio of the radial position of an electron to the beam radius.

V Potential by which the electrons have been accelerated, volts.

v Velocity, meters/second.

Φ Angular dimension of beam cross section.

x, y Rectangular co-ordinates in a cross-section plane through the beam.

z Position of an electron in the axial direction.

ω A convenient variable equal to dr_e/dt.

Subscript 1 indicates the value of the variable at a particular cross-section plane.

Electrical Characteristics of Power Transistors*

ALLEN NUSSBAUM†

Summary—**The following electrical characteristics of a group of power transistors whose collector dissipation is in the region of 20 watts have been measured: the small-signal current amplification as a function of emitter current, the cutoff frequency as a function of emitter current and collector voltage, and the decrease of current amplification with increasing frequency. All these measurements were made in both common-base and common-emitter configuration. The results were compared with the theory developed for low-power transistors by Shockley, Rittner, Webster, and others, and it was found that for the frequency-dependent parameters, the agreement was quite good, but that differences exist in the case of those which are a function of current. This leads to the conclusion that further work must be done on the theory of junction power transistors.**

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Introduction

A GENERAL description of p-n-p diffused junction power transistors having a collector dissipation in the region of 20 watts has been given by Roka, Buck, and Reiland.[1] Constructional features are shown in Fig. 1 (next page). Fig. 2 (next page) gives the characteristic curves for common-emitter operation of three typical production units, representing respectively the approximate upper and lower limits of quality and an average unit. The curves shown were taken on a recording potentiometer, and some of the high current

[1] E. G. Roka, R. E. Buck, G. W. Reiland, "Developmental germanium power transistors," Proc. I.R.E., vol. 42, pp. 1247–1250; August, 1954.

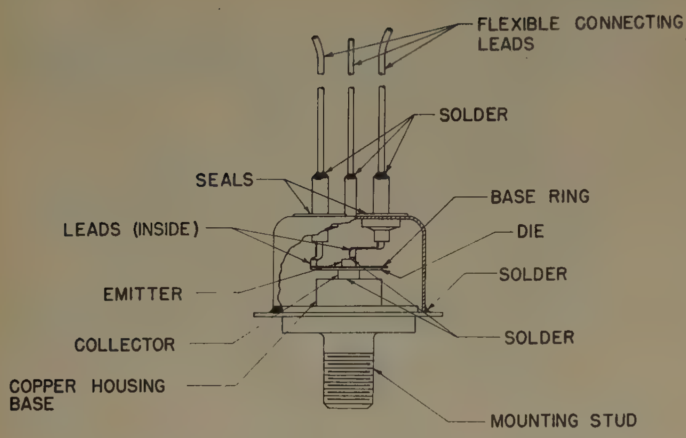


Fig. 1—Cross-sectional drawing of a power transistor.

curves show a slight negative slope due to heating effects. Oscillographic curves of the same units do not show this behavior. In the following sections, we shall give a detailed report on some of the electrical properties of these transistors, and attempt to correlate these results with theory.

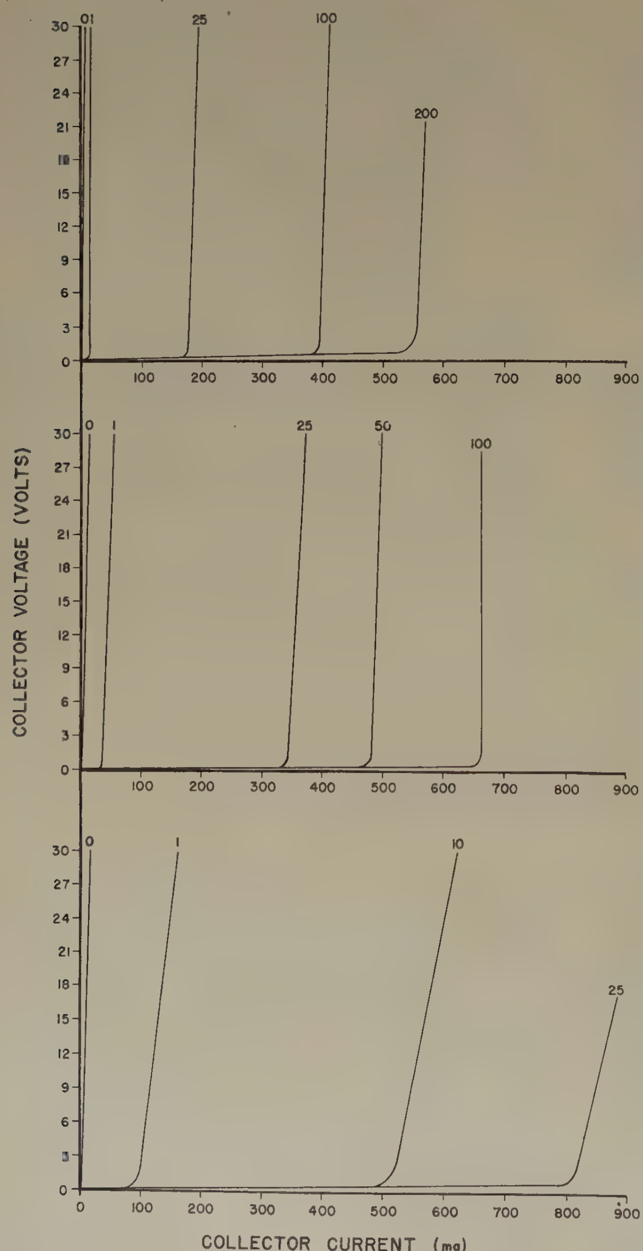
Current Amplification as a Function of Emitter Current

Experimental Results

The current amplification α of a transistor for common base operation is defined

$$\alpha = \partial I_c / \partial I_e) V_c, \tag{1}$$

where I_e is the emitter current, I_c is the collector cur-



TRANSISTOR NO. 1063			
V_b	I_b	I_c	V_c
(FLOATING)		.63ma	30V
.035V .085	0 ma	.51 3.8	2 30
.19 .20	1	12 19	2 30
1.3 1.3	25	185 200	2 30
3.0 3.2	100	400 415	2 30
4.6 4.8	200	555 575	2 21

TRANSISTOR NO. 355			
V_b	I_b	I_c	V_c
(FLOATING)		.63ma	30V
.04V .15	0ma	.18 14.5	2 30
.25 .25	1	38 50	2 30
1.05 1.10	25	340 370	2 30
1.4 1.6	50	485 500	2 30
2.1 2.4	100	660 655	2 30

TRANSISTOR NO. 1055			
V_b	I_b	I_c	V_c
(FLOATING)		.42ma	30V
.036V .115	0ma 0ma	.69 18.0	2 30
.28 .28	1 1	102 160	2 30
1.0 1.05	10 10	530 620	2 30
1.8 1.9	25 25	815 885	2 17
4.2	100	1500	2

NOTE: IDENTIFYING NUMBERS ON CURVES DENOTE BASE CURRENT IN ma.

Fig. 2—Common-emitter characteristic curves of three representative power transistors.

rent, and V_c is the collector voltage (see List of Symbols). For common-emitter operation the current amplification b is

$$b = \partial I_c/\partial I_b)V_c, \tag{2}$$

where I_b is the base current. These parameters were measured by superposing small ac signals on the dc quantities I_e (or I_b) and I_c and measuring their ratio.

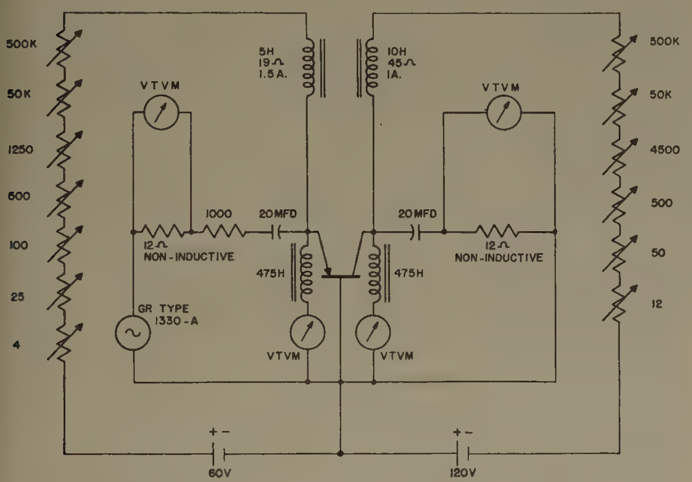


Fig. 3—Circuit diagram of current-amplification measuring apparatus.

The circuit used is shown in Fig. 3. Various direct-reading methods reported in the literature[2] were not usable because of the lower impedances and higher currents associated with power transistors. Chokes to handle the currents involved were specially wound and the bias sources consisted of high-current dry batteries connected in series. The frequency of measurement was 400 cycles per second. The results for common-base operation are plotted in Fig. 4, and for common-emitter operation in Fig. 5.

Theoretical Dependence of Current Amplification on Emitter Current

When the p-n-p transistor is considered to be a device of rectangular shape with base region of width w, with diffusion of carriers restricted to one dimension, and with the field in the base being small enough so that carrier transport is due entirely to diffusion, then the common-base current amplification can be expressed[3]

$$\alpha = \beta\gamma, \tag{3}$$

where

$$\beta = \operatorname{sech}(w/L_b), \tag{4}$$

and

$$\gamma = \frac{1}{1 + (D_n/D_p)(n_e/p_b)(L_b/L_e)\tanh(w/L_b)}. \tag{5}$$

In these equations β is the diminuation factor due to volume recombination, γ is the injection efficiency; D_n and D_p are the diffusion constants of electrons in the emitter and holes in the base, respectively; n_e and p_b are the corresponding equilibrium concentrations; and L_e and L_b are the corresponding diffusion lengths. We also have $L_e^2 = D_p\tau_b$, where τ_b is the lifetime of holes in the base.

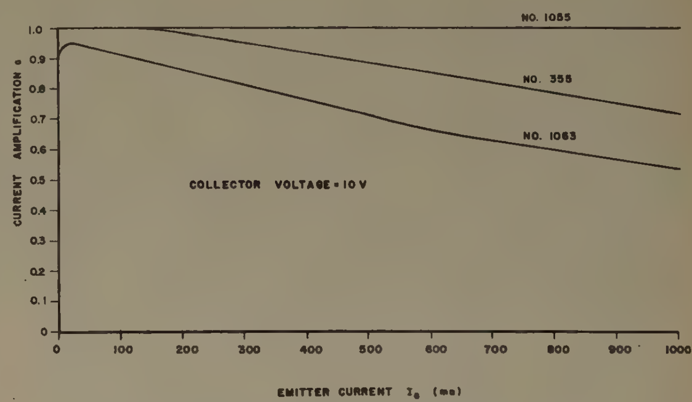


Fig. 4—Emitter current vs current amplification for common-base operation.

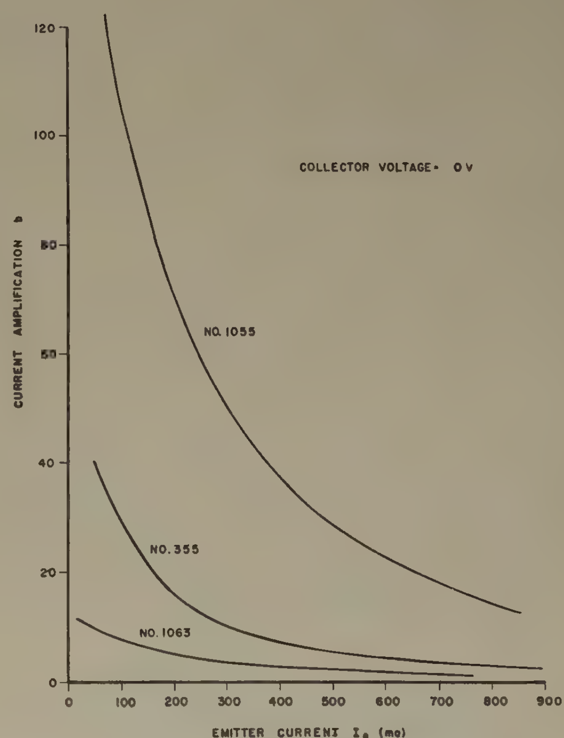


Fig. 5—Emitter current vs current amplification for common-emitter operation.

The common emitter-current amplification is

$$b = \alpha/(1 - \alpha), \tag{6}$$

and since the argument w/L_b of the hyperbolic functions is known to be small enough to permit series expansion, (6) may be converted into

[2] See, for example, R. F. Shea, "Principles of Transistor Circuits," John Wiley and Sons, Ch. 22, New York; 1953.

[3] E. L. Steele, "Theory of alpha for p-n-p diffused junction transistors," Proc. I.R.E., vol. 40, pp. 1424–1428; November, 1952.

$$
\begin{aligned}
1/b &= 1 - \beta\gamma \\
&= 1 - (1 + w^2/2L_b^2)^{-1}(1 + D_n n_e w/D_p p_b L_e)^{-1} \quad (7)\\
&= (w^2/2D_p\tau_b) + (D_n n_e w/D_p p_b L_e),
\end{aligned}
$$

where only first order terms have been retained. If the effects of surface recombination are considered, Rittner[4] and Webster[5] have shown that an additional term must be included in (7):

$$1/b = (w^2/2D_p\tau_b)+(D_n n_e w/D_p p_b L_e) + (sA_s w/D_p A_e), \quad (8)$$

where s is the surface recombination velocity, A_s is the effective surface area for recombination, and A_e is the area of the conduction path which is approximately equal to the emitter area. All of the parameters on the right of (8) with the exception of τ_b are constants of the transistor and independent of I_e. It has been shown by Shockley and Read[6] that τ_b should vary with injected carrier density δn in the following fashion:

$$\tau_b = \tau_0(1 + x\delta n)/(1 + y\delta n), \quad (9)$$

where τ_0 is the lifetime in the limit of no injected carriers and where we can compute $x=9.66\times10^{-16}/\text{cm}^3$ and $y=2.58\times10^{-15}/\text{cm}^3$ from data for lifetime vs impurity concentration given by Hall.[7] The resistivity of the starting material used in these transistors is 4.5 ohm-cm and from this it was determined that $\tau_0=295$ μsec with τ_b decreasing to a minimum of 110 μsec as the injected current rises. This should cause a change in the volume recombination term $w^2/2D_p\tau_b$ by a factor of 3. However, both Webster[5] and Rittner[4] have computed that for receiving-type transistors, both the volume-recombination and injection-efficiency terms are about 0.1 the surface-recombination term. Hence, the variation in τ_b has little effect on $1/b$ and (8) predicts that the current-amplification is independent of emitter current. Webster has shown this to be untrue for both p-n-p and n-p-n receiving-type transistors and Fig. 5 shows similar behavior in power transistors. To consider the reason for this dependence, it should be recalled that in the derivation of (3), it was assumed that the electric field in the base region of the transistor appears only across the junctions and can be neglected in the remainder of the n-type area. By considering the effect of this field, Webster has shown that (8) must be modified to

$$
\begin{aligned}
1/b = {} & [(w^2/2D_p\tau_b) + (\sigma_b w/\sigma_e L_e)](1 + Z)\\
& +(swA_s/D_p A_e)g(Z), \quad (10)
\end{aligned}
$$

where

$$Z = (w\mu_e/D_p A_e \sigma_b)I_e \quad (11)$$

and

$$g(Z) = \frac{1 + p_1/N_d}{1 + 2p_1/N_d}; \quad (12a)$$

$$Z = (2p_1/N_d) - \ln(1 + p_1/N_d). \quad (12b)$$

In the above equations μ_e is the electron mobility, p_1 the hole concentration in the base at the emitter junction, N_d the donor ion concentration, and the emitter and base conductivities σ_e and σ_b, respectively, enter through replacing the ratio of diffusion constants by use of the Einstein relation and the definition of conductivity in terms of mobility. Eqs. (12a) and (12b) show that the function $g(Z)$ decreases rapidly with Z, so that the surface-recombination term in (10) produces a rise in b with current, whereas the other two terms cause a decrease, and the curve of b vs I_e shows a maximum. The physical picture is that at low currents the volume-recombination rate is low, the injection-efficiency is high, but a large number of holes are being lost by diffusion to the surface. Surface recombination is presumed to occur in an annular ring surrounding the emitter and of width approximately equal to w. As the current (and the field) increases, there is less diffusion to the surface, causing b to rise, but at the same time there is a fall in b due to the other two factors. By using Lagrange's multipliers and considering (12a) and (12b) as a constraint, the maximum of b as a function of I_e is found to occur when

$$(2g-1)^2 g = [(w^2/2D_p\tau_b)+(\sigma_b w/\sigma_e L_e)]/(swA_s/D_p A_e). \quad (13)$$

This relation will be applied subsequently.

Comparison with Experiment

The physical constants and the computed values of the various terms in (10) for the receiving-type transistors of Webster and for one of the transistors of Fig. 2 are listed in Table I. The curve for transistor No. 1055 as given in Fig. 5 was used in conjunction with (10) for obtaining the values of $\sigma_e L_e$ and sA_s given in the table. This particular transistor was chosen since it has an α of approximately 1.0 over the entire I_e range, this being one of the assumptions involved in Webster's

TABLE I

Parameter	RCA	MH No. 1055
Emitter diameter, d	0.015 inch	0.130 inch
Collector diameter	0.045 inch	0.150 inch
Base width, w	4.8×10^{-3} cm	6.35×10^{-3} cm
Emitter area, A_e	1.1×10^{-3} cm²	8.55×10^{-2} cm²
Base resistivity, $1/\sigma_b$	2.22 ohm-cm	4.5 ohm-cm
Lifetime, τ_b	500 μsec	295 μsec
$w^2/2D_p\tau_b$	0.0006	0.0016
sA_s (from best fit)	0.147 cm³/sec	5.0 cm³/sec
$sA_s w/D_p A$	0.014	0.0062
$\sigma_e L_e$ (from best fit)	1.55/ohm	1.0/ohm
$\sigma_b w/\sigma_e L_e$	0.0014	0.0014
Z	800 I_e(I_e in amp)	27.3 I_e(I_e in amp)
Current for which b is a maximum, computed from (13)	1.2 ma	0.73 ma
s (assuming $A_s=w\pi d$)	245 cm/sec	760 cm/sec

[4] E. S. Rittner, "Extension of the theory of the junction transistor," *Phys. Rev.*, vol. 94, pp. 1161–1171; June, 1954.

[5] W. M. Webster, "On the variation of junction-transistor current amplification factor with emitter current," Proc. I.R.E., vol. 42, pp. 914–920; June, 1954.

[6] W. Shockley and W. T. Read, "Statistics of the Recombinations of Holes and Electrons," *Phys. Rev.*, vol. 87, pp. 835–840; September, 1952.

[7] R. N. Hall, "Electron-hole recombination in germanium," *Phys. Rev.*, vol. 87, p. 387; July, 1952.

theory. The principal differences between the RCA and Honeywell transistors are the size of the junctions and the magnitude of the currents carried. Because of the insensitivity of the measuring circuit at low currents it was not possible to confirm the existence of the maximum in b predicted by (13). We would expect, however, that if this maximum exists, it should occur at currents much higher than 1 ma, for when the two types of transistor are carrying the same current, the current density (and hence the field) in the power transistor should be smaller. This means that the effect of surface recombination is larger and the maximum should be shifted to the right in the I_e vs b curve. It must be concluded, therefore, that the first-order approximations which give such reasonable agreement for the smaller units do not hold for high-current transistors. The more rigorous one-dimensional theory of Rittner also predicts a curve of the same shape as Webster's. Rittner has extended his work to three dimensions in a most elegant manner, and shows that for transistors of the size he considers, the one-dimensional theory is a reasonable approximation. In view of the above discussion, it is felt at this point that in order to bring theory and experiment into line for power transistors, further work will have to be done with the three-dimensional theory with regard to converting the formidable solutions of the equations into a form suitable for numerical computation.

Frequency-Dependent Characteristics

Common Base Configuration

Returning to the conditions under which (3) is valid, it has been shown by Steele[3] that if the injected hole density varies sinusoidally with angular frequency ω, then this equation is still applicable provided we replace w/L_b with zw/L_b, where $z^2=(1+i\omega\tau_b)$. For low currents, the approximation is made that $\alpha=1.0$, and all of the frequency variation of α is contained in β, where

$$\beta = \operatorname{sech}(zw/L_b), \tag{14}$$

or

$$\begin{aligned}\beta &= [1 - (w^2/2L_b^2) - (i\omega\tau_b w^2/2L_b^2)]^{-1} \\ &= [1 - i\omega\tau_b w^2/2L_b^2]^{-1}.\end{aligned} \tag{15}$$

(The use of the series expansion for a function with a complex argument can be justified by computing the absolute value and incorporating the condition that $w/L_b \ll 1$.) The common-base cutoff frequency f_{ca} is defined as the frequency for which the magnitude squared of β falls to one half of its low-frequency value ($f=0$), or

$$\beta(f_{ca})\beta^*(f_{ca})/\beta^2(0) = 1/2. \tag{16}$$

It can be shown that if the real part of β is independent of frequency, then f_{ca} can alternatively be defined as the frequency for which the real and imaginary parts of β are equal. Using this latter condition on (15),

$$f_{ca} = D_p/w^2\pi. \tag{17}$$

On the other hand, substituting β as given by (14) into the definition (16) yields, after taking the square root of the complex number z and expanding the hyperbolic secant of the resulting root,

$$\cosh 2x + \cos 2y = 4, \tag{18}$$

where x and y are defined by

$$\begin{aligned}x + iy = wz/L_b &= (w/L_b)(1 + \omega_{ca}^2\tau_b^2)^{1/4} \\ &\cdot [\cos(\theta/2) + i \sin(\theta/2)],\end{aligned} \tag{19}$$

and

$$\theta = \arctan \omega_{ca}\tau_b$$
$$f_{ca} = 2\pi\omega_{ca}. \tag{20}$$

The common-base cutoff frequency is plotted in Fig. 6 as a function of emitter current and collector voltage. For these frequencies we have $\omega_{ca}\tau_b \gg 1$, so that $\theta \approx \pi/2$, $x=y$, and (18) has a solution of $x=1.1$, or, by (19),

$$1.1 = (\omega_{ca}\tau_b w^2/2L_b^2)^{1/2}$$

or

$$f_{ca} = 1.22\, D_p/\pi w^2. \tag{21}$$

This relation, given by Rittner[4] and Pritchard,[8] predicts an f_{ca} 22 per cent higher than (17). For w as given in Table I,

$$f_{ca} = 1.22 \times 44/\pi \times (6.35 \times 10^{-3})^2 = 422 \text{ kc/sec.}$$

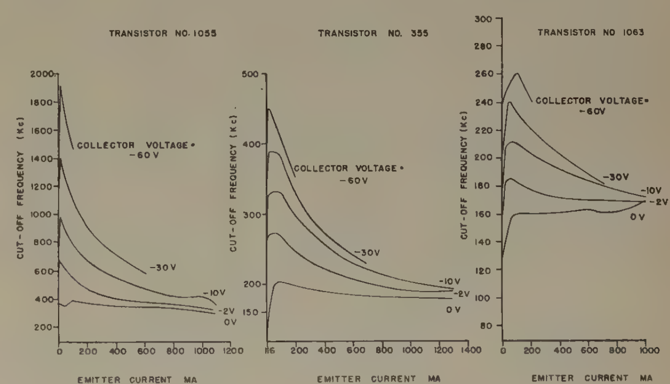


Fig. 6—Common-base cutoff frequency as a function of emitter current and collector voltage.

This is reasonable for transistor No. 355 at low currents. As I_e increases, Rittner predicts that f_{ca} should rise to twice its low-current value and then decline somewhat, due to a fall-off in injection efficiency. This general behavior is seen in Fig. 6, but here again we are faced with computational difficulties for quantitative comparisons.

The frequency dependence of α (or β) can be obtained by substituting (17) into (15),

$$\alpha = \beta = (1 - if/f_{ca})^{-1}, \tag{22}$$

[8] R. L. Pritchard, "Frequency variations of current-amplification factor for junction transistors," Proc. I.R.E., vol. 40, pp. 1476–1481; November, 1952.

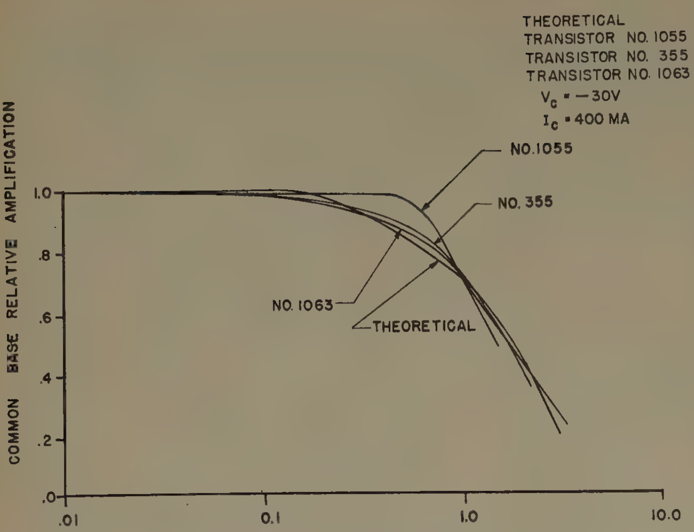


Fig. 7—Common-base relative frequency vs relative amplification.

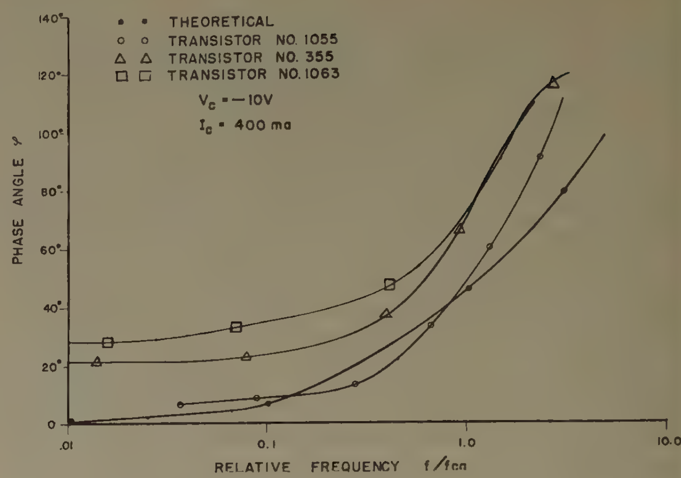


Fig. 8—Phase angle vs relative frequency for common-base operation.

or in polar form,

$$\alpha = A(f)e^{i\phi}, \tag{23}$$

where

$$A(f) = [1 + (f/f_{ca})^2]^{-1/2} \tag{24a}$$

$$\phi = \arc \tan f/f_{ca}. \tag{24b}$$

By (24a)

$$A(f)/A(0) = [1 + (f/f_{ca})^2]^{-1/2} \tag{25}$$

As shown by Pritchard,[8] (25) and (24a) mean that the relative amplitude $A(f)/A(0)$ and the phase ϕ of α are universal functions of the relative frequency f/f_{ca}. A comparison of the plot of (24a) with measured values is given in Fig. 7. The collector current and voltage used were selected for convenience. Little dependence on these quantities was observed. The agreement for transistors No. 355 and 1063 is quite good. The deviation for No. 1055 might be due to a frequency variation in γ suggested by Pritchard.[9] The measurement of phase angle was found to be difficult using standard Lissajous[10] methods, so that it was necessary to compute ϕ in the following manner:

$$b = Be^{i\Delta}$$
$$= Ae^{i\phi}/(1 - Ae^{i\phi}),$$

where B is the amplitude of b and Δ the phase angle. Then

$$B^2 = bb^* = A^2(1 + A^2 - 2A \cos \phi),$$

or

$$\cos \phi = (1 + A^2/B^2)/2A \tag{26}$$

Fig. 8 shows experimental values of ϕ as computed from (26) and the theoretical values computed from (25).

[9] R. L. Pritchard, "Frequency variations of junction-transistor parameters," Proc. I.R.E., vol. 42, pp. 786–799; May, 1954.

[10] F. E. Terman and J. M. Pettit, "Electronic Measurements," McGraw-Hill Book Co., pp. 267–268, New York; 1952.

Common Emitter Configuration

For convenience in derivation of equations, we will interchange the order of treatment in this section. First we will get an approximate expression for f_{cb}, the cutoff frequency for this configuration. From (6) and (15), assuming as usual that α is not a function of frequency and $w/L_b \ll 1$,

$$b = [\cosh (zw/L_b) - 1]^{-1}$$
$$= [(z^2w^2/2L_b^2) + z^4w^4/4!L_b^4)]^{-1}. \tag{27}$$

Equating real and imaginary parts

$$f_{cb} = 1/2\pi\tau_b \tag{28}$$

so that under the assumption that volume recombination is of major importance at high injection levels, f_{cb} should rise with an increase in I_e. A linear rise is seen in Fig. 9 (opposite). The change in lifetime from 295 to 110 μsec should give corresponding values of f_{cb} from 0.55 to 1.5 kc/sec, so that (28) is not a very good approximation formula.

To find the dependence of b on f_{cb}, put (28) into (27) to get

$$b = [(w^2/2D_p\tau_b)(1 + if/f_{cb})]^{-1} \tag{29}$$

or

$$B(f)/B(0) = [1 + (f/f_{cb})^2]^{-1/2}, \tag{30}$$

which is of the same form as (25). This curve and the experimental results are shown in Fig. 10 (opposite page).

The cutoff frequency definition as exemplified by (16), when applied to the combination of (7) and (14), yields, using the same symbols as (19),

$$b^2(0)/2 = [\cosh (zw/L_b) - 1]^{-1}[\cosh (z^*w/L_b) - 1]^{-1}$$

or

$$\cosh 2x + \cos 2x - 4 \cosh x \cos x + 2 = 0, \tag{31}$$

where we have dropped a term $4/b^2(0)$. The solution is $x = 0.743$, or

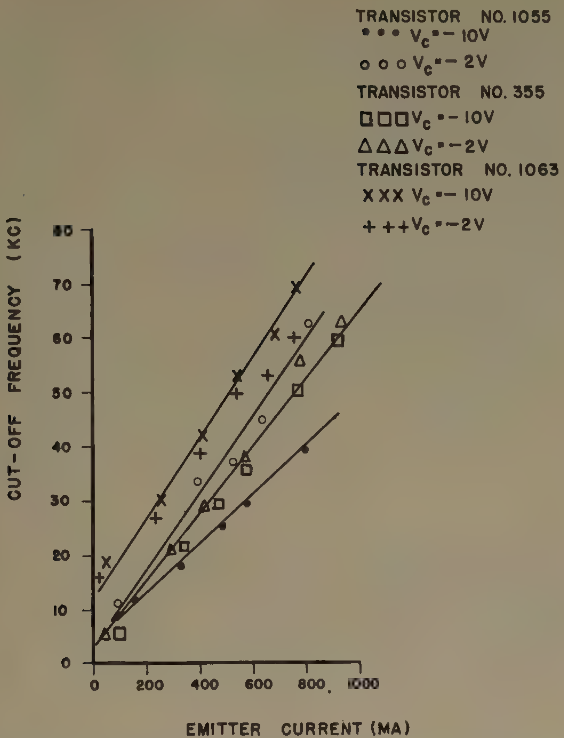


Fig. 9—Common-emitter cutoff frequency as a function of emitter current and collector voltage.

$$f_{cb} = 0.549\, D_p/\pi w^2 \tag{32}$$

and

$$f_{cb} = 190 \text{ kc/sec.}$$

This is not a good approximation, either, but is in the opposite direction to (28).

A third method of estimating f_{cb} has been given by Thomas.[11] By (22),

$$\alpha(f) = \alpha(0)/(1 - if/f_{ca}),$$

so that

$$b(f) = \alpha(f)\{1 - \alpha(f)\}$$

$$= \frac{\alpha(0)}{\{1-\alpha(0)\}\left[1 - \dfrac{if}{f_{ca}\{1-\alpha(0)\}}\right]}. \tag{33}$$

By (29),

$$b(f) = b(0)/(1 + if/f_{cb}), \tag{34}$$

hence

$$f_{cb} = \{1 - \alpha(0)\} f_{ca}. \tag{35}$$

Estimating an error of about 1 per cent in the measurement of α, this means that f_{cb} should be of the order of 0.01 f_{ca} for Nos. 1055 and 355 at low currents. This prediction is quite reasonable. The agreement for No. 1063 is similar. However, (35) does not predict the rise in f_{cb} as f_{ca} falls.

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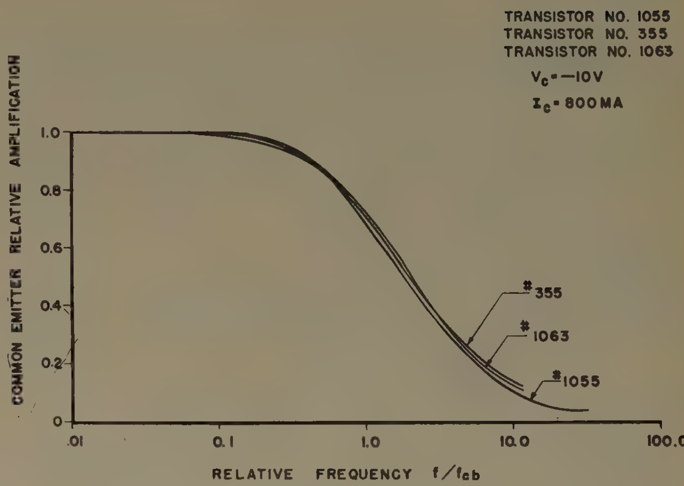

Fig. 10—Common-emitter relative frequency vs relative amplification.

Conclusions

It is apparent from the data presented that, in the case of power transistors, successful devices were manufactured before their theory of operation was too well understood. It is felt that numerical calculations based on Rittner's work hold out the most hope for deriving expressions connecting current-amplification and cutoff frequency with current density in transistors carrying large currents and having large junctions. It is also planned to test these results by making similar studies on units of about 60 watts dissipation, now in the process of development.

Acknowledgment

The writer would like to express his appreciation to Finn J. Larsen and Van W. Bearinger, who provided continuous encouragement and support of this project. A special note of thanks is due John R. Crane, who constructed all the measuring circuits and performed the tedious task of taking the data.

List of Symbols

- A = amplitude of α
- A_e = emitter area
- A_s = surface recombination area
- B = amplitude of b
- b = current amplification for common-emitter operation
- D_n = diffusion constant of electrons
- D_p = diffusion constant of holes
- d = emitter diameter
- f_{ca} = common-base cutoff frequency
- f_{cb} = common-emitter cutoff frequency
- $g(Z) = (1+p_1/N_d)/(1+2p_1/N_d)$
- I_b = base current (dc)
- I_c = collector current (dc)
- I_e = emitter current (dc)
- L_e = diffusion length in emitter
- L_b = diffusion length in base
- N_d = donor ion concentration

n_e = equilibrium electron concentration in emitter
p_1 = hole concentration in base at emitter junction
p_b = equilibrium hole concentration in base
s = surface recombination velocity
V_c = collector voltage
w = base width (distance from emitter to collector junctions)
$x+iy = wz/L_b$
$z = (1+i\omega\tau_b)^{1/2}$
$Z = w\mu_e/D_pA\sigma_b = (2p_1/N_d) - ln(1+p_1/N_d)$
α = current amplification for common base operation
β = diminuation factor
γ = injection efficiency
Δ = phase angle of b
δn = injected-carrier density
θ = arc tan $\omega_{ca}\tau_b$
μ_e = electron mobility
σ_b = base conductivity
σ_e = emitter conductivity
τ_0 = minority carrier lifetime with no injected carrier density
τ_b = lifetime of holes in base
ϕ = phase angle of α
ω = angular frequency of injected hole current
$\omega_{ca} = 2\pi f_{ca}$

Further Bounds Existing on the Transient Responses of Various Types of Networks*

A. H. ZEMANIAN†, ASSOCIATE, IRE

Summary—**Five new theorems stating bounds on the transient response of certain types of networks are derived and illustrated. The first one states that any system function, whose real part on the positive real-frequency axis decreases monotonically as frequency increases, cannot have an overshoot in its step response greater than eighteen per cent, nor can its rise time, from the time that the input step is applied to the time that the response first crosses the final value line, be less than $1.22(r-K)C$ where r is the final value of the step response and K and $1/C$ are the constant term and the coefficient of the $1/s$ term in the inverse power-series expansion of the system function. Similarly, the other four theorems show that when the system function is appropriately restricted, the transient response is bounded. These restrictions include never-negative or never-positive conditions on the real or imaginary parts of the system function along the real-frequency axis, while the resulting bounds are determined by certain coefficients occurring in the power-series expansions of the system function.**

Introduction

Many empirical rules exist which relate characteristics of the frequency response of systems to characteristics of their transient response.[1–4] Furthermore, the frequency responses of many types of networks, such as passive driving point impedances or minimum phase networks, are restrained in various ways.[5]

In a previous paper,[6] analogous bounds have been shown to exist on the transient response when the frequency response of the system is appropriately restricted. Moreover, these relations are exact ones and not approximate rules of thumb. In this paper theorems stating new exact relations of this type are proven and illustrated.

Theorems

Only fixed, lumped, linear, and stable networks will be considered in the forthcoming discussion where the system function, $Z(s)$, has no poles on the real-frequency axis or in the right-half s-plane. Since this system function is analytic at $s=0$ and $s=\infty$, it may be expanded around these two points into the following two infinite series, respectively:

$$Z(s) = r + k_1s + k_2s^2 + k_3s^3 + \cdots; \quad |s| \leqq q,$$

$$Z(s) = K + \frac{1}{Cs} + \frac{K_2}{s^2} + \frac{K_3}{s^3} + \cdots; \quad |s| \geqq p.$$

As these expansions will be used only in the vicinity of $s=0$ and $s=\infty$, let p be a number greater than the distance from the origin to the pole of $Z(s)$ furthest from the origin, and let q be a number less than the distance from the origin to the pole nearest the origin. In this paper, the first expression will be called the power-series expansion of $Z(s)$ and the second, the inverse power-series expansion of $Z(s)$. As usual all input functions are assumed applied at time, $t=0$, so that all responses are zero for negative values of time.

The first theorem states that any network, whose system function has a real part along the real-frequency

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axis $R(\omega)$, which decreases monotonically as ω increases from zero toward infinity, will have a unit step response, $A(t)$, whose percentage overshoot will be less than 18 per cent and whose rise time from the time the step input is applied to the time the response crosses the final value line will be greater than $1.22(r-K)C$. This situation is illustrated in Fig. 1 for the case where $K=0$. When the derivative of $R(\omega)$ is excluded from the positive region shown shaded in Fig. 1(a), the unit step response is excluded from those regions which are shaded in Fig. 1(b). An example also is indicated wherein the real part of the system function and the corresponding unit step response are realizable by a shunt-peaked filter whose poles are at $s=-1\pm j1$ and whose zero is at $s=-2$. The statement of this theorem and its proof follow.

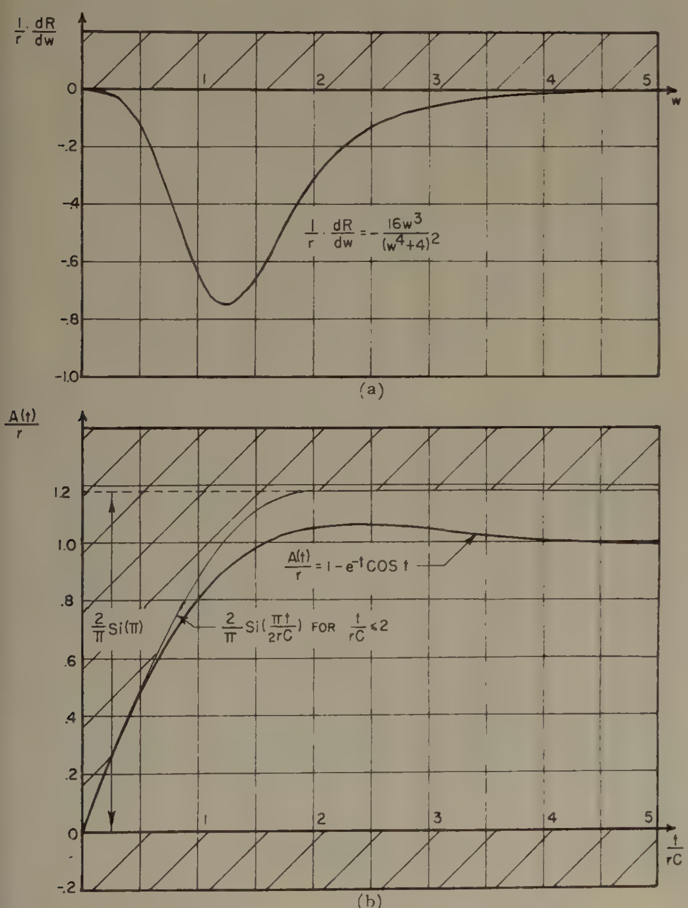

Fig. 1—Prohibited regions of Theorem 1 for the case $K=0$. (a) The region from which dR/dw is excluded by the hypothesis of Theorem 1. (b) The regions into which the corresponding $A(t)$ cannot enter.

Theorem 1

If $Z(s)$ is such that $(dR/d\omega)\leqq 0$ for $0\leqq\omega$, then for $0\leqq t\leqq 2(r-K)C$:

$$K \leqq A(t) \leqq K+\frac{2}{\pi}(r-K)\,Si\left[\pi\frac{t}{2(r-K)C}\right]$$

and for $2(r-K)C\leqq t$:

$$K \leqq A(t) \leqq K+\frac{2}{\pi}(r-K)\,Si(\pi) < K+1.18(r-K).$$

Proof: Consider the Fourier transform which expresses the unit step response in terms of the real part of the system function.

$$A(t)=\frac{2}{\pi}\int_0^\infty R(\omega)\frac{\sin\omega t}{\omega}d\omega \quad (1)$$

$$=\frac{2}{\pi}t\int_0^{\pi/t}R(\omega)\frac{\sin\omega t}{\omega t}d\omega + \frac{2}{\pi}t\int_{\pi/t}^{\infty}R(\omega)\frac{\sin\omega t}{\omega t}d\omega \quad (2)$$

$$=G(t)+H(t).$$

Denoting the second term in expression (2) by $H(t)$ and integrating it by parts, the following may be obtained:

$$H(t)=K-\frac{2}{\pi}Si(\pi)R\left(\frac{\pi}{t}\right)+\frac{2}{\pi}\int_{\pi/t}^{\infty}\left(-\frac{dR}{d\omega}\right)Si(\omega t)\,d\omega.$$

Since $(dR/d\omega)\leqq 0$ and $Si(\omega t)\leqq Si(\pi)$,

$$H(t)\leqq K-\frac{2}{\pi}Si(\pi)R\left(\frac{\pi}{t}\right)+\frac{2}{\pi}Si(\pi)\int_{\pi/t}^{\infty}\left(-\frac{dR}{d\omega}\right)d\omega$$

$$H(t)\leqq K\left[1-\frac{2}{\pi}Si(\pi)\right]. \quad (3)$$

Now a bound will be obtained on the first term of expression (2) which will be denoted by the symbol, $G(t)$. Within the interval $0\leqq\omega\leqq(\pi/t)$, the function $(\sin \omega t/\omega t)$ is positive and monotonically decreasing. Therefore, of all the positive functions, $R(\omega)$, such that $(dR/d\omega)\leqq 0$, the rectangular pulse function, $R_m(\omega)$, is the one which will yield the greatest value of $G(t)$.

$$\begin{aligned}R_m(\omega)&=r \text{ for } |\omega|\leqq\omega_c\\ &=K \text{ for } \omega_c\leqq|\omega|\end{aligned} \quad (4)$$

where $K<r$. Integrating $[Z(s)-K]$ around the right half s-plane and making use of the fact that there are no poles in the right half s-plane nor on the $j\omega$ axis, the following expression may be obtained:[5]

$$\frac{2}{\pi}\int_0^\infty[R(\omega)-K]d\omega=\frac{1}{C}\cdot$$

Applying this to $Rm(\omega)$ and solving for ω_c,

$$\omega_c=\frac{\pi}{2(r-K)C}\cdot$$

Thus, when $(\pi/t)\geqq\omega_c$ so that $t\leqq 2(r-K)C$,

$$G(t)\leqq\frac{2}{\pi}t\int_0^{\omega_c}r\frac{\sin\omega t}{\omega t}d\omega+\frac{2}{\pi}t\int_{\omega_c}^{\pi/t}K\frac{\sin\omega t}{\omega t}d\omega$$

$$\leqq\frac{2}{\pi}[r-K]\,Si(\omega_c t)+\frac{2}{\pi}K\,Si(\pi). \quad (5)$$

Adding (3) and (5),

$$A(t) \leqq K + \frac{2}{\pi}[r - K]\, Si\left[\pi \frac{t}{2(r-K)C}\right]. \qquad (6)$$

Similarly, when $(\pi/t) \leqq \omega_c$, so that $t \geqq 2(r-K)C$:

$$G(t) \leqq \frac{2}{\pi} Si(\pi) r.$$

Therefore

$$A(t) \leqq K + \frac{2}{\pi} Si(\pi)[r - K]. \qquad (7)$$

Expressions (6) and (7) constitute the upper bounds of this theorem. The fact that the unit step response corresponding to $R_m(\omega)$ equals this bound for $0 \leqq t \leqq 2(r-K)C$ demonstrates that these are the lowest possible bounds of this form.

To obtain the lower bound, integrate (1) by parts:

$$A(t) = K + \frac{2}{\pi}\int_0^\infty \left(-\frac{dR}{d\omega}\right) Si(\omega t) d\omega.$$

The fact that both terms in the integrand of this expression are never negative establishes this lower bound. It is the greatest possible constant one, since $A(0) = K$. This completes the proof.

When $(dR/d\omega) \geqq 0$ for $0 \leqq \omega$, so that $K > r$, it may be shown in exactly the same way that for $0 \leqq t \leqq 2(r-K)C$:

$$K - \frac{2}{\pi}[K - r]\, Si\left[\pi \frac{t}{2(r-K)C}\right] \leqq A(t) \leqq K,$$

and for $2(r-K)C \leqq t$:

$$K - \frac{2}{\pi}[K - r]\, Si(\pi) \leqq A(t) \leqq K.$$

In these expressions, C is a negative quantity.

The second theorem improves the result of Theorem 2 of reference 6. It shows that any network, whose system function has a real part along the positive real-frequency axis $R(\omega)$ which decreases monotonically to zero as ω increases from zero toward infinity, will have a unit impulse response, $W(t)$, which is bounded by the functions shown in Fig. 2. The example is obtained from the same system function used in illustrating Theorem 1.

Theorem 2

If $Z(s)$ is such that $K = 0$ and $(dR/d\omega) \leqq 0$ for $0 \leqq \omega$, then for $0 \leqq t \leqq rC$:

$$|W(t)| \leqq \frac{1}{C}\,\frac{\sin\dfrac{\pi t}{2rC}}{\dfrac{\pi t}{2rC}}$$

and for $rC \leqq t$:

$$|W(t)| \leqq \frac{2r}{\pi t}.$$

Proof: The impulse response may be written in terms of the real part of the system function as follows:

$$W(t) = \frac{2}{\pi}\int_0^{\pi/2t} R(\omega)\cos\omega t d\omega + \frac{2}{\pi}\int_{\pi/2t}^{\infty} R(\omega)\cos\omega t d\omega \qquad (8)$$
$$= J(t) + L(t).$$

Denoting the second term in (8) by $L(t)$, integrating it by parts and using the fact that $\sin \omega t$ is always less than one, it can be shown that $L(t)$ is a negative quantity. Therefore, $W(t) \leqq J(t)$.

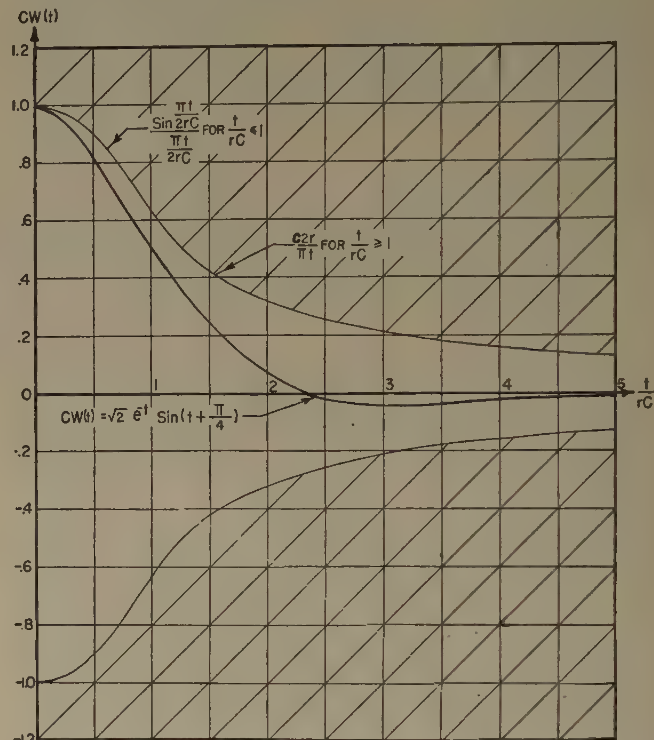

Fig. 2—Prohibited regions of the conclusion of Theorem 2.

Of all the positive functions, $R(\omega)$, such that $(dR/d\omega) \leqq 0$, the rectangular pulse function given by (4) is the one which will yield the greatest value of $J(t)$ and thereby the upper bounds. Thus, for $(\pi/2t) \leqq \omega_c$ so that $t \geqq rC$:

$$W(t) \leqq \frac{2}{\pi}\int_0^{\pi/2t} r\cos\omega t d\omega = \frac{2r}{\pi t};$$

and for $(\pi/2t) \geqq \omega_c$ so that $t \leqq rC$

$$W(t) \leqq \frac{2}{\pi}\int_0^{\omega_c} r\cos\omega t d\omega = \frac{1}{C}\,\frac{\sin\dfrac{\pi t}{2rC}}{\dfrac{\pi t}{2rC}}.$$

The lower bounds are obtained in the same way except that use is made of the fact that $\sin\omega t \geqq -1$. Consideration of the impulse response corresponding to $R_m(\omega)$ demonstrates that the upper bounds are the lowest possible ones of this form.

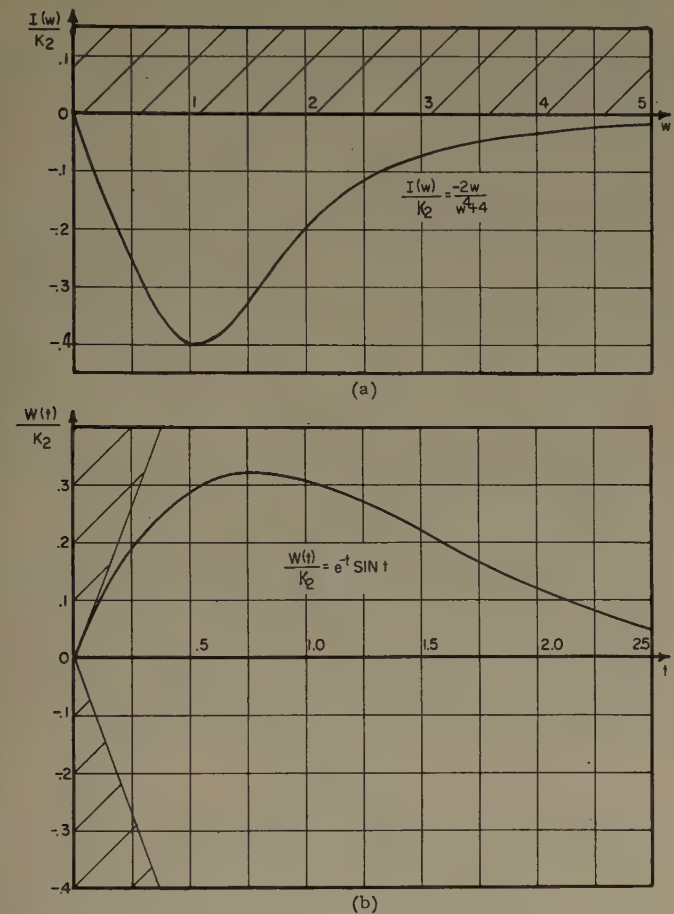

Fig. 3—Prohibited regions of Theorem 3. (a) The region from which $I(w)$ is excluded by the hypothesis of Theorem 3. (b) The regions into which the corresponding $W(t)$ cannot enter.

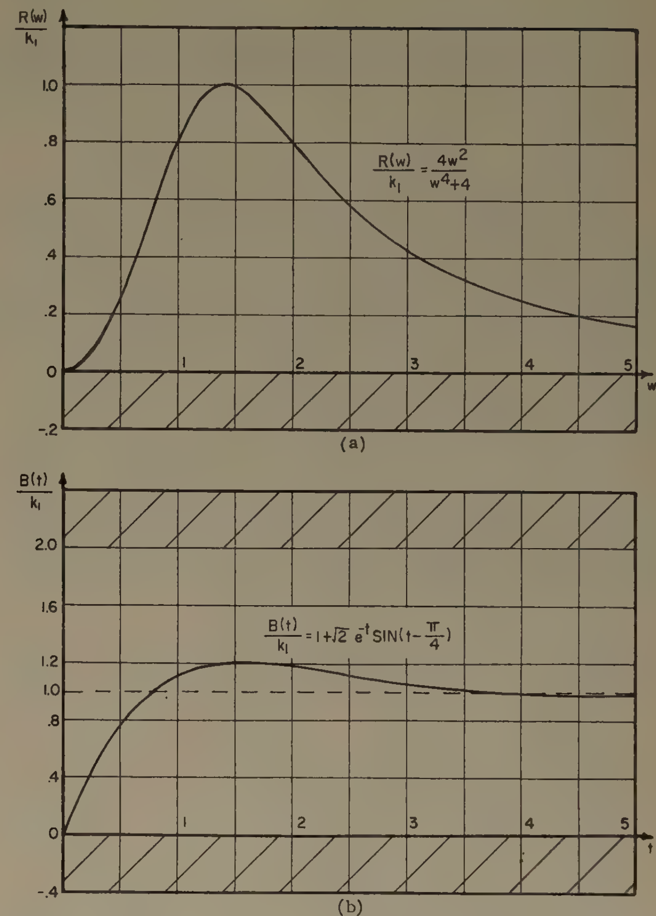

Fig. 4—Prohibited regions of Theorem 4. (a) The region from which $R(w)$ is excluded by the hypothesis of Theorem 4. (b) The regions into which the corresponding $B(t)$ cannot enter.

The third theorem states that any network, whose system function has two more poles than zeros and has a never-positive imaginary part along the positive real-frequency axis, $I(\omega)$, will have a unit impulse response, $W(t)$, whose magnitude is bounded by K_2t. This, of course, means that the magnitude of the unit step response is bounded by $(K_2/2)t^2$. The shaded portions of Fig. 3 indicate the regions from which the imaginary part and the corresponding impulse response are excluded by the hypothesis and conclusion of this theorem, respectively. Again an example is given. The functions shown are obtained from a system function having only two poles at $s=-1\pm j1$.

Theorem 3

If $Z(s)$ is such that $K=1/C=0$ and $I(\omega)\leqq 0$ for $0\leqq\omega$, then $|W(t)|\leqq K_2t$.

Proof: The usual argument is applied to the Fourier sine transform which relates the unit impulse response to the imaginary part of the system function.

$$W(t) = -\frac{2}{\pi}\int_0^\infty I(\omega)\sin\omega t\,d\omega. \tag{9}$$

Since $I(\omega)$ is never-positive and $|\sin\omega t|\leqq\omega t$,

$$|W(t)| \leqq -\frac{2}{\pi}t\int_0^\infty \omega I(\omega)d\omega. \tag{10}$$

Integrating $sZ(s)$ around the right-half s-plane and using the fact that there are no poles on or within this contour (if the path of integration is indented to the right at the origin to avoid the simple pole occurring there), a value for the integral in (10) may be found.[5]

$$\frac{2}{\pi}\int_0^\infty \omega I(\omega)d\omega = -K_2.$$

Thus:

$$|W(t)| \leqq K_2t.$$

That this bound is the best possible linear one which passes through the origin can be shown by applying the initial value theorem to the derivative of $W(t)$.

$$\lim_{t\to 0}\frac{dW}{dt} = \lim_{s\to\infty} s^2Z(s) = K_2;$$

therefore $W(t)$ is tangent to K_2t at the origin. This completes the proof.

The fourth theorem states that when the system function has a simple zero at the origin and has a never-negative real part on the real-frequency axis, the corresponding unit ramp response, $B(t)$, will never be less than zero nor greater than $2k_1$. This unit ramp function is defined as being zero for negative values of time, t, and equal to t for positive values of time. Fig. 4 il-

lustrates the restricted regions of the hypothesis and conclusion while the example is obtained from a system function having two poles at $s=-1\pm j1$ and a zero at the origin.

Theorem 4

If $Z(s)$ is such that $r=0$ and $R(\omega)\geqq 0$ for all ω, then $0\leqq B(t)\leqq 2k_1$.

Similarly, the last theorem states that, when the system function has a double zero at the origin and its imaginary part is never-negative along the positive real-frequency axis, the corresponding unit parabolic response, $C(t)$, will never be less than zero nor greater than $4k_2$. The unit parabolic function is defined as being zero for negative values of time, t, and equal to t^2 for positive values of time. Again these restricted regions on $I(\omega)$ and $C(t)$ are shown in Fig. 5. The example is obtained from a system function having two poles at $s=1\pm j1$ and a double zero at the origin.

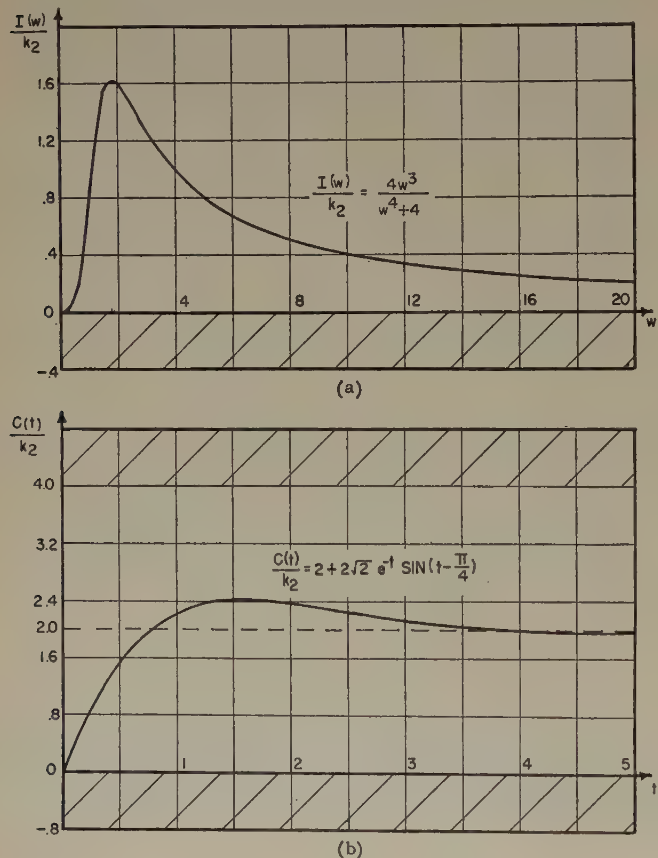

Fig. 5—Prohibited regions of Theorem 5. (a) The region from which $I(w)$ is excluded by the hypothesis of Theorem 5. (b) The regions into which the corresponding $C(t)$ cannot enter.

Theorem 5

If $Z(s)$ is such that $r=k_1=0$ and $I(\omega)\geqq 0$ for $0\leqq\omega$, then $0\leqq C(t)\leqq 4K_2$.

Theorem 4 can be obtained from Theorem 3 of reference 6 by considering $Z(s)/s$ as the system function. In this case, $-(R(\omega)/\omega)$ replaces $I(\omega)$: k_1 replaces r; and $B(t)$ replaces $A(t)$. Similarly, Theorem 5 can be obtained from the same theorem of reference 6 by considering $2(Z(s)/s^2$ as the system function and letting $-2(I(\omega)/\omega^2)$ replace $I(\omega)$, $2k_2$ replace r, and $C(t)$ replace $A(t)$. This process of finding bounds on responses to inputs of the form t^n for system functions with multiple zeros of nth order at the origin cannot be continued, for neither $R(\omega)$ nor $I(\omega)$ can be never-negative or never-positive for these higher order cases where $r=k_1=k_2=0$. On the other hand, the examples given for each theorem demonstrate that the conditions of these theorems are realizable. To show that the bounds of these last two theorems are the best possible constant ones, one need only consider an $R(\omega)$ or $I(\omega)$ which is a very narrow pulse at some positive and negative value of ω and zero elsewhere. Corresponding responses will approach their bounds at overshoots and undershoots.

Conclusion

The results of this paper demonstrate the possibility of reflecting bounds which might occur on the responses of systems in the frequency domain into bounds which hold on their responses in the time domain. These effects may be interpreted as practical limitations on the rise times or overshoots of the responses of general types of networks.

Appendix

- $A(t)$ The response to a unit step function applied at time, $t=0$.
- $B(t)$ The response to the input function, t, applied at time, $t=0$.
- C The reciprocal of the coefficient of $1/s$ in the power-series expansion of $Z(s)$.
- $C(t)$ The response to the input function, t^2, applied at time, $t=0$.
- $G(t)$ The first term in (2).
- $H(t)$ The second term in expression (2).
- $I(\omega)$ The imaginary part of a system function for real frequencies.
- $J(t)$ The first term in (8).
- K The constant term in the inverse power-series expansion of $Z(s)$.
- K_2 The coefficient of $1/s^2$ in the inverse power series expansion of $Z(s)$.
- k_1 The coefficient of s in the power-series expansion of $Z(s)$.
- k_2 The coefficient of s^2 in the power-series expansion of $Z(s)$.
- $L(t)$ The second term in (8).
- $R(\omega)$ The real part of a system function for real frequencies.
- $R_m(\omega)$ Rectangular pulse function defined by (4).
- r The constant term in the power-series expansion of $Z(s)$.
- $Si(\omega t)$ $\int_0^{\omega t}(\sin x/x)dx$.
- s A complex variable.
- $W(t)$ The response to a unit impulse function applied at time, $t=0$.
- ω (2π) (frequency) = angular frequency.
- $Z(s)$ A system function.

Magnetic Focusing of Electron Beams*

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Summary—**Electron trajectories for a pencil beam are studied for both uniform and periodic magnetic fields. General equations are given which apply to shielded and unshielded cathodes and all intermediate cases. The balance conditions which yield minimum ripple solutions are presented as a function of magnetic field. It is shown that these conditions require that considerable magnetic flux thread the cathode for magnetic fields in excess of the Brillouin value**

Introduction

THE DYNAMICS of an electron beam guided by an axial field is well known, and in particular several special cases have been studied in great detail. Probably the most familiar amongst these special cases is "Brillouin" focusing for which the cathode is completely shielded from the magnetic field.[1] A pseudo-type of Brillouin focusing in which a periodic magnetic field was employed has recently been described in the literature.[2,3] Some work has also been done on "confined" systems in which the cathode is immersed in a long uniform magnetic field.

It is the purpose of this paper to show the simple relationship which exists between these various systems, and several new systems which have not been discussed heretofore in the literature. Considered here for the first time are cases intermediate between a completely shielded and an unshielded cathode both for uniform and periodic magnetic fields. The optimum conditions of operation are derived and simple expressions for the electron trajectories are given. A semi-shielded cathode possesses the advantage of inhibiting the spread of the beam in the gun region due to thermal velocities while still retaining some of the magnetic efficiency of complete shielding. For a periodic system the weight-saving feature of alternately poled permanent magnets can be realized while the low noise properties of a "confined" system are presumably retained.

Dynamics

In the general case of an unshielded cathode, where an arbitrary amount of magnetic flux is allowed to thread the electron beam, Bush's theorem becomes

$$\dot{\theta} = \frac{\eta}{2\pi r^2}(N_2 - N_1); \tag{1}$$

where $N_2 = B\pi r^2$, $N_1 = B_c\pi r_c^2$, B is the longitudinal magnetic field at r, B_c is the longitudinal magnetic field at the cathode, and r_c is the radius of the electron beam as it leaves the cathode. It is assumed that the initial $\dot{\theta}$ is zero (at the cathode) and that there is no radial variation of the magnetic field.

Combining the three forces (magnetic, centrifugal, and space-charge repulsion) acting on an edge electron of a round beam in an axially symmetrical magnetic field, one obtains

$$\frac{d^2r}{dt^2} + r\left(\frac{\eta^2B^2}{4}\right) - r\left(\frac{\eta^2B_c^2}{4}\right)\left(\frac{r_c}{r}\right)^4 - \frac{r_0^2\omega_p^2}{2r} = 0; \tag{2}$$

where

$$\omega_p^2 = \frac{\eta\rho}{\epsilon},$$

$$\rho = \frac{I}{\sqrt{2\eta v_0}\pi r_0^2},$$

and r_0 is the reference radius from which ρ is calculated (see Fig. 1).

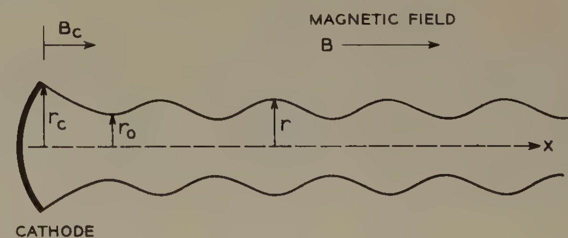

Fig. 1—Schematic diagram of the beam (r_0 generally, but not necessarily, equals r_{min} from the gun).

Note that (2) is quite general and makes no assumptions as to variation of the longitudinal magnetic field.

As a practical matter most electron guns (for traveling-wave tubes) are convergent, which leads to r_c (cathode radius) being larger than r_0, the reference radius of the beam. However, there is no loss of generality if r_c is set equal to r_0, since B_c can still be adjusted at will. It is only the product $B_cr_c^2$ which is significant, so either may be fixed. Letting $r_c = r_0$, (2) becomes

$$\ddot{r} + r\left(\frac{\eta^2B^2}{4}\right) - \frac{r_0^4}{r^3}\left(\frac{\eta^2B_c^2}{4}\right) - \frac{r_0^2}{r}\left(\frac{\omega_p^2}{2}\right) = 0.$$

Further simplification results from letting $y = r/r_0$:

$$\ddot{y} + y\left(\frac{\eta^2B^2}{4}\right) - \frac{1}{y^3}\left(\frac{\eta^2B_c^2}{4}\right) - \frac{1}{y}\left(\frac{\omega_p^2}{2}\right) = 0. \tag{3}$$

This is a non-linear differential equation about which little is known unless certain restrictions are imposed on

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† Bell Telephone Labs, Inc., Murray Hill, N. J.

[1] C. C. Wang, "Electron beams in symmetric fields," Proc. I.R.E., vol. 38, pp. 135–147; February, 1950.

[2] J. T. Mendel, C. F. Quate, and W. H. Yocum, "Electron beam focusing with periodic permanent magnets," Proc. I.R.E., vol. 42, pp. 800–810; May, 1954.

[3] A. M. Clogston and H. Heffner, "Focusing of an electron beam by periodic fields," *Jour. Appl. Phys.*, vol. 25, pp. 436–447; April, 1954.

the variable. Since we are interested primarily in stable electron beams the important solutions occur when y is close to unity. Therefore, let $y=1+\delta$. For $\delta \ll 1$,

$$\frac{1}{(1+\delta)^3} \cong 1 - 3\delta,$$

and

$$\frac{1}{(1+\delta)} \cong 1 - \delta.$$

Eq. (3) becomes

$$\ddot{\delta} + \left(\frac{\eta^2 B^2}{4} + \frac{3\eta^2 B_c^2}{4} + \frac{\omega_p^2}{2}\right)\delta + \left[\frac{\eta^2 B^2}{4} - \frac{\omega_p^2}{2} - \frac{\eta^2 B_c^2}{4}\right] = 0. \tag{4}$$

For magnetic fields which have a uniform region B_0, the following notation can be used:

$$\omega_L = \tfrac{1}{2}\eta B_0$$

$$K = \left(\frac{B_c}{B_0}\right)^2 = \left(\frac{N_1}{N_2}\right)^2.$$

Substituting the above,

$$\ddot{\delta} + \left[\omega_L^2(1+3K) + \frac{\omega_p^2}{2}\right]\delta + \left[\omega_L^2(1-K) - \frac{\omega_p^2}{2}\right] = 0. \tag{5}$$

Here K is just the square of the fractional percentage of the flux threading the cathode. For complete shielding of cathode (Brillouin focusing) $K=0$. For no shielding at all (confined flow) $K=1$. The solution of (5) is

$$\delta = C_1 \sin\left[\omega_L^2(1+3K) + \frac{\omega_p^2}{2}\right]^{1/2} t + C_2 \cos\left[\omega_L^2(1+3K) + \frac{\omega_p^2}{2}\right]^{1/2} t - \frac{\left[\omega_L^2(1-K) - \frac{\omega_p^2}{2}\right]}{\left[\omega_L^2(1+3K) + \frac{\omega_p^2}{2}\right]}, \tag{6}$$

where

$$C_1 = \frac{\frac{d\delta}{dt}(0)}{\left[\omega_L^2(1+3K) + \frac{\omega_p^2}{2}\right]^{1/2}} \tag{7}$$

and

$$C_2 = \delta(0) + \frac{\left[\omega_L^2(1-K) - \frac{\omega_p^2}{2}\right]}{\left[\omega_L^2(1-3K) + \frac{\omega_p^2}{2}\right]}. \tag{8}$$

Arbitrarily we can say that the beginning of the uniform magnetic field occurs when $t=0$.

In most practical cases it is desirable to keep the perturbations of the beam edge to a minimum. To accomplish this several alternatives avail themselves. If we let $d\delta/dt=0$ at $t=0$ (start the beam parallel in the uniform field) then $C_1=0$. If we further stipulate that at $t=0$, $\delta=0$; that is, at the beginning of the uniform field $r=r_0$, then the minimum ripple solution is obtained when

$$\frac{\left[\omega_L^2(1-K) - \frac{\omega_p^2}{2}\right]}{\left[\omega_L^2(1+3K) + \frac{\omega_p^2}{2}\right]} = 0,$$

or

$$\omega_L^2(1-K) - \frac{\omega_p^2}{2} = 0,$$

and

$$K = 1 - \frac{\left(\frac{\omega_p^2}{2}\right)}{\omega_L^2}.$$

For Brillouin-type focusing,

$$(\omega_L^2) = \left(\frac{\omega_p^2}{2}\right)$$

and $K=0$. In practice it is generally necessary to increase the magnetic field above the Brillouin value by at least a factor of 1.5 to take care of transverse thermal velocities and lens aberrations. For $\omega_L = 1.5\ \omega_{L\text{Brillouin}}$,

$$K = 0.55 = \left(\frac{B_c}{B_0}\right)^2$$

$$B_c = 0.75\ B_0,$$

or 75 per cent of the flux should thread the cathode for a minimum ripple solution if the magnetic field is increased 50 per cent above the Brillouin value. For higher magnetic fields this percentage goes up. For the case of larger magnetic fields it should be noted that the ripple due to a finite input derivative is less [see (7)].

If we remove the requirement that $\delta=0$ at $t=0$, then referring to (8) we can make the constant $C_2=0$ if

$$\delta(0) = -\frac{\left[\omega_L^2(1-K) - \frac{\omega_p^2}{2}\right]}{\left[\omega_L^2(1+3K) + \frac{\omega_p^2}{2}\right]}.$$

That is, we start the beam in the uniform field at a radius smaller than r_0 (assuming a field greater than Brillouin) and completely shield the cathode. This brings us right back to Brillouin flow with a smaller beam radius. It is argued (by some people) that this is a

better approach since it yields a smaller diameter beam which presumably will go through a smaller diameter tube. In a highly convergent system this cannot always be done since here there are limits beyond which it is impossible to reduce the size of the beam. Also, controlling the beam in this way (reducing the diameter to compensate for the increased magnetic field) leads to a nonuniform current distribution over the cross section and larger space-charge forces.

By the simple adjustment of the parameter K the previous equations can be made to describe all uniform systems lying between complete shielding and complete immersion of the cathode in the magnetic field. Thus the comparison of particular cases is greatly simplified.

The periodic counterpart of uniform Brillouin focusing has already been shown in the literature,[2,3] but as yet periodic focusing with flux threading the cathode has not been treated. It would seem desirable to develop this in the above manner using the parameter K.

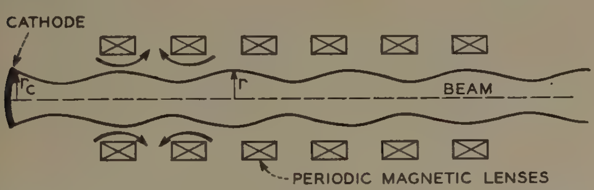

Fig. 2—Schematic diagram of the beam in a periodic magnetic field. Here L equals twice the distance between lens centers.

Periodic Fields

For sinusoidally varying fields let $B=B_0 \cos x$ (longitudinal field) where $x=\omega t$ and $\omega=2\pi u_0/L$, $\omega_L=1/2\eta B_0$, as in Fig. 2. Rewriting (4) with a periodic field, we have

$$\ddot{\delta}+\left(\frac{\omega_L^2}{2}(1+\cos 2x)+3K\omega_L^2+\frac{\omega_p^2}{2}\right)\delta + \left[\frac{\omega_L^2}{2}(1+\cos 2x)-\frac{\omega_p^2}{2}-K\omega_L^2\right]=0. \qquad (9)$$

Substituting

$$\frac{d^2\delta}{dt^2}=\omega^2\frac{d^2\delta}{dx^2}$$

and

$$\alpha=\frac{1}{2}\left(\frac{\omega_L}{\omega}\right)^2 \quad \text{magnetic field parameter}$$

$$\beta=\frac{1}{2}\left(\frac{\omega_p}{\omega}\right)^2 \quad \text{space charge parameter,}$$

$$\delta''+[\alpha(1+\cos 2x)+6\alpha K+\beta]\delta + [\alpha(1+\cos 2x)-\beta-2K\alpha]=0, \qquad (10)$$

where the primes indicate differentiation with respect to x. Eq. (10) is readily solved if the term $(\alpha \cos 2x)\delta$ is dropped. Such an approximation can be justified by comparison to an exact solution obtained from the analog computer, and by physical arguments. Since we are restricting ourselves to cases where the flux threads, the cathode, K will be of the order of 1/2 in most practical cases. Consequently we are ignoring only about 20 per cent of the second term of (10). With this approximation we can write

$$\delta=C_1\cos(\alpha+6\alpha K+\beta)^{1/2}x+C_2\sin(\alpha+6\alpha K+\beta)^{1/2}x + \frac{\beta+2K\alpha-\alpha}{(\alpha+6\alpha K+\beta)}+\frac{\alpha\cos 2x}{4-(\alpha+6\alpha K+\beta)}. \qquad (11)$$

To minimize beam ripple let

$$\beta+2K\alpha-\alpha=0$$

$$K=\frac{\alpha-\beta}{2\alpha}. \qquad (12)$$

For $\alpha=\beta$ (periodic Brillouin flow), K will be zero and the cathode completely shielded. For α (magnetic field) far in excess of the Brillouin value K approaches 1/2, or 70 per cent of the peak, magnetic field should thread the cathode for minimum ripple. If the balance conditions of (12) are maintained, (11) becomes

$$\delta=C_1\cos(4\alpha-2\beta)^{1/2}x+C_2\sin(4\alpha-2\beta)^{1/2}x + \frac{\alpha\cos 2x}{4-(4\alpha-2\beta)}. \qquad (13)$$

If

$$\delta(0)=\frac{\alpha}{4-(4\alpha-2\beta)^{1/2}},$$

then $C_1=0$ and

$$C_2=\frac{\delta'(0)}{(4\alpha-2\beta)}.$$

For the above conditions,

$$\delta=\frac{\delta'(0)}{(4\alpha-2\beta)^{1/2}}\sin(4\alpha-2\beta)^{1/2}x+\frac{\alpha\cos 2x}{4-(4\alpha-2\beta)}; \qquad (14)$$

$\delta'(0)$ is the initial derivative of the beam profile as it enters the peak of the periodic field. In more familiar terms

$$\delta'(0)=\frac{L}{2\pi r_0}\frac{dr}{dz}(0).$$

In practice the first term of (14) generally predominates, so to obtain a smooth beam it behooves one to pay close attention to the problem of starting the beam out parallel.

Beam Stability

The approximate forms of (3) are intended to serve only as a guide to the magnitude and period of the small ripple solutions and are not intended as criteria of stability. The power series expansion of $1/y^3$ and $1/y$ assumed values of y close to unity, and therefore the equations in δ are suspect even before regions of instability are approached. It is necessary to locate the

unstable regions (divergent flow) from (3) (with $B=B_0 \cos x$) since this equation contains no assumptions concerning the value of y. Thus we have

$$y'' + \alpha(1 + \cos 2x)y - \frac{2\alpha K}{y^3} - \frac{\beta}{y} = 0. \qquad (15)$$

The first two terms of (15) comprise a form of Mathieu's equation which has received extensive treatment in the literature.[2,3] It was found experimentally[2] and with the analog computer that the term $-\beta/y$ does not seem to alter the location of the first stop band (unstable region) which seems to be wholly dependent upon α. One could further argue that in view of the above, the term $-(2\alpha K/y^3)$ would contribute negligibly to the stability of the solution. Judging from the computer solutions for a variety of conditions this indeed seems to be the case. To predict behavior in the higher order pass and stop bands one must be more circumspect since both $-(B/y)$ and $-(2\alpha K/y^3)$ contribute singularities at $y=0$ which are absent in Mathieu's equation. It could be argued that the $-(\beta/y)$ term is not actually a singularity in a physical situation because one can never obtain a point focus of an electron beam. However, the $-(2\alpha K/y^3)$ term is a real singularity regardless of the aberrations in the system. As long as some flux threads the cathode no electrons can cross the axis (this can be seen from simple energy considerations). Just how this effects the higher order pass bands is not known. The experimental results of the shielded gun case ($K=0$) indicate that the higher order pass and stop bands are not significantly different from those of Mathieu's equation alone.

There seems to be no simple way of including the $1/y$ and $1/y^3$ terms in a stability criterion, which gives further inducement for the above assumptions.

Comparison of the Shielded and Unshielded Cases

It is difficult to classify categorically one system as superior to another unless a specific application was the basis for comparison. One can, however, enumerate several advantages of both systems.

Shielded Cathode

1. A shielded cathode will focus the greatest amount of space charge with a given magnetic field.
2. Only electrostatic forces exist in the cathode-anode region since the gun is shielded from the magnetic field.
3. With a fixed period of magnetic field a greater amount of space charge can be focused in the first pass band.

Unshielded Cathode

1. A balance condition with respect to the various forces acting on the beam can be maintained regardless of how much the magnetic field is varied merely by altering the flux threading the cathode. This means that magnetic fields far in excess of the Brillouin value can be employed while still maintaining a minimum ripple solution. This may be advantageous in a low-noise device.
2. In some instances it may be easier to construct an unshielded gun since the proximity of the cathode to the magnetic circuit is not so critical. (In the shielded case the magnets must begin at the beam minimum which is generally close to the cathode). Also no magnetic materials such as shield cans need be used near the gun.
3. Thermal spread is inhibited in the gun region.

Square-Wave Magnetic Fields

The necessity of getting the rf energy on and off a traveling-wave tube focused with periodic magnetic fields creates serious electrical problems, especially when the period is quite small. Any disturbance of the magnetic field strength or period is likely to initiate a large ripple on the beam. Therefore, it is desirable to incorporate the waveguide or other matching system within the periodic structure. If the individual lenses are shorter than the waveguide width then sometning must be done to either the matching section or the magnetic structure. To obviate the necessity of making these changes (which are difficult) it has been suggested that a long period be used where the magnetic field is made to approximate a square wave rather than a sinusoid.

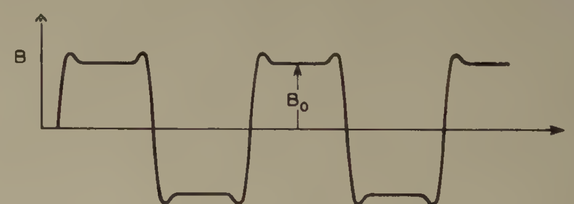

Fig. 3—The variation of magnetic field with distance along the axis for a peaked square-wave system.

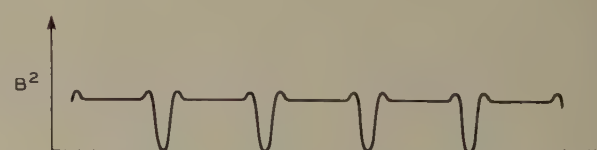

Fig. 4—The variation of magnetic focusing *force* with distance (since force is proportional to B^2).

In order to circumvent the disturbance experienced by an electron in going between two lenses, J. R. Pierce has proposed peaking the field in the manner shown in Figs. 3 and 4. As can be seen from (3), it is the square of the magnetic field which determines the inward focusing force on the electrons. If the peaking of the magnetic

field is properly adjusted the average impulse that an electron receives in crossing a transition between two lenses can be made zero. That is,

$$B^2 < \text{average} > = B_0^2$$

For this condition one might expect an electron to traverse the transition with a minimum disturbance.

If in addition to the above requirement the peaks and troughs of Fig. 4 are adjusted so that the fundamental Fourier component is absent, then the lowest order stop band will have been eliminated. In principle this argument can be extended to the second, third, and higher harmonics. As demonstrated by Pipes[4] for any periodic waveform of the form in Fig. 4, the first unstable region of Hill's equation occurs when the natural period of the wave solution is twice the period of the lowest order harmonic of the square of the magnetic field as shown in Fig. 5. We are again assuming that the terms in $1/y$ and $1/y^3$ do not materially alter the stability criterion.

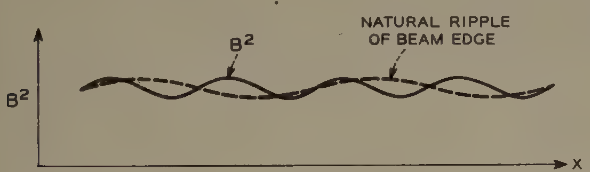

Fig. 5—The conditions for the first unstable region, where the ripple period equals twice the period of B^2.

The process of limiting the harmonic content of the magnetic field and thereby removing stop bands has certain physical limitations in that the resulting wave shape must be producible with a physical structure. Cut-and-try procedures based on the above indicate that the best structure seems to be one in which the transition is made as short as possible and the impulse made to be zero. Any further refinement seems to complicate the system disproportionately.

Traveling-wave tubes are relatively short in terms of the number of magnetic lenses necessary to focus the beam. If the operation is such that the beam radius is diverging exponentially, it may not appreciably effect the current transmission if the divergence is slow enough. Thus a stop band does not always imply no transmission. Many uniform solenoids have small periodic perturbations of field strength which in theory produce stop bands although these are not detected by observing percentage transmission of beam current.

In order to obtain quantitative measurements, (3) with a B of the form shown in Fig. 3 was solved on the analog computer for various magnitudes of B and B_c (field strength at the cathode). The particular form of B was chosen because it resembled a physical structure under investigation.

[4] L. A. Pipes, "Matrix solution of equations of the Mathieu-Hill type," *Jour. Appl. Phys.*, vol. 24, pp. 902–910; July, 1953.

J. W. Sullivan of these Laboratories, during the course of his investigations of focusing a low noise traveling-wave tube, analyzed the above wave in terms of its Fourier expansion. The results are given in Fig. 6. Although the first and third harmonics are present to some extent the instability was associated with the fourth harmonic (that is, the period of the beam profile just prior to the unstable point was approximately twice that of the fourth harmonic). This is not unusual since the amplitude of the fourth harmonic is quite large compared to the rest. There are two explanations for the apparent absence of instability for the first and third harmonics. First, the width of a stop band is proportional to the amplitude of the corresponding sinusoidal component in the magnetic field. The width may have been so small we inadvertently missed it. Second, the degree of divergence (the rate of growth of the beam profile) is also proportional to the magnitude of the sinusoidal component, and because this was small for the first and third harmonics, they might have gone undetected since only about 10 periods were followed on the computer (for the case studied this represents a length of travel for the beam greater than any practical traveling-wave tube).

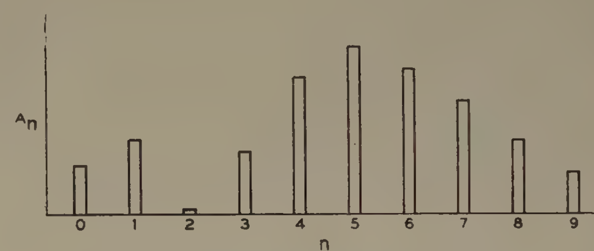

Fig. 6—Fourier components of the peaked magnetic field $B = B_0 + \sum An \cos nx$.

Conclusion

To summarize it can be said that some of the foregoing relationships appeared as somewhat of a surprise to the author. The fact that the balance conditions require a high percentage of flux threading the cathode for a relatively small increase in magnetic field above the Brillouin value is not generally known. For these conditions it is possible in principle to obtain a beam with a more uniform current density and smaller perturbations of the beam edge than is the case for Brillouin focusing. The presence of some magnetic field at the cathode tends to inhibit the motion due to thermal velocities and therefore can be very useful especially with high-density beams. It is interesting that these principles hold equally well for uniform or periodic fields.

Acknowledgment

The author owes a great deal to C. C. Cutler of these Laboratories for suggesting the approach to the problem used in this paper.

The Mitron—An Interdigital Voltage-Tunable Magnetron*

J. A. BOYD†, MEMBER, IRE

Summary—**This paper describes a voltage-tunable magnetron which can be electronically tuned in the frequency range 1,500 to 3,500 mc. This tube delivers a useful power output of approximately 200 mw and, when used as a local oscillator in a microwave receiver, gives a noise figure 3 db greater than is obtained when a reflex klystron is used with the same receiver. It is shown that temperature-limited operation of the cathode is necessary for coherent operation of the tube and that the operating voltage is greater than that predicted by the Hartree relation. The relation of tube operation and circuit parameters is discussed.**

Introduction

This paper describes the design, construction and operation of a microwave generator that can be used in many of the applications where electronically-tunable power is needed. This generator consists of an external-cavity interdigital magnetron, which has been called a mitron, and its associated cavity structure. The mitron can be electronically tuned in the frequency range of 1,500 to 3,500 mc and delivers a useful power output in the order of 200 mw.

The frequency of operation of the mitron can be made dependent on the anode potential for certain operating conditions. This mode of operation has been called "voltage tuning," and is defined here as the variation in the frequency of the output signal from a magnetron as the anode potential is varied (when the dc anode current, magnetic field, and load impedance are maintained constant).[1] Voltage-tuning was first observed by P. H. Peters and D. A. Wilbur.[2] Using conventional power magnetrons with special low-Q or nonresonant circuits, they obtained voltage-tuning in the frequency range 500 to 1,000 mc. Dr. J. S. Needle[3] obtained voltage-tuning in the frequency range of 2,000 to 3,000 mc, using a bar-and-vane type magnetron structure. The power output from this tube was in the order of 10 to 30 microwatts of coherent signal power.

Significant results which are discussed in this paper are: (1) the operation of voltage-tunable magnetrons, with useful power output, has been obtained in the frequency range 1,500 to 3,500 mc, as compared to the 500 to 1,000 mc range at the General Electric Company; (2) much higher power has been obtained than was obtained by Needle; (3) much data have been obtained which contribute to knowledge of the operation of voltage-tunable magnetrons.

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† Dept. Elec. Engrg., Univ. of Michigan, Ann Arbor, Mich.
[1] H. W. Welch, Jr., "Prediction of traveling-wave magnetron frequency characteristics: frequency pushing and voltage tuning," Proc. I.R.E., vol. 41, pp. 1631–1653; November, 1953.
[2] D. A. Wilbur, R. B. Nelson, P. H. Peters, A. J. King, and L. R. Koller, "C. W. Magnetron Research," Final Rept., Contract No. 2-36-039 sc32279, Res. Lab., Gen. Elec. Co.; April, 1950.
[3] J. S. Needle, "The Insertion Magnetron," Univ. of Michigan Electron Tube Lab. Tech. Rep. No. 11; August, 1951.

Design of Mitron and Cavity

Design of Cavity

In the design of this magnetron and its cavity, two factors were considered essential if a satisfactory voltage-tunable magnetron was to be developed. These factors were: (1) low anode-to-anode capacitance; and (2) A high-impedance external cavity, the impedance being a slowly varying function of frequency (by high impedance, we mean in the order of 100 ohms).

The interdigital anode structure is adaptable to mounting in a waveguide structure and was selected for this reason. It also has wide-range tuning possibilities and is mechanically simple.

The ridge waveguide offers a possibility for increasing the circuit impedance over that of a rectangular waveguide for a given spacing between anode supports. The ridge waveguide is characterized by its lowered cut-off frequency and lowered impedance (compared with a rectangular guide of equal dimensions), and by a wide bandwidth free from higher-mode interference.

A section of standard S-band rectangular waveguide was used with a ridge 0.84 inch high and 1.0 inch wide. This structure has a characteristic impedance at infinite frequency of 160 ohms and a cut-off frequency of 1,250 mc. Since it is desirable to have a coaxial output from the mitron, a tapered-ridge waveguide-to-coaxial-line junction was used to terminate each end of the ridge waveguide. This arrangement is shown in Fig. 1, facing page. Junctions used are based on a design given by Cohn.[4] They have an upper cutoff frequency of approximately 4,800 mc; therefore, when used with a ridge waveguide give a very wide-band structure.

Design of Mitron

The Interaction Space: In the design of the interaction space for the mitron, there were two factors, the anode-to-anode capacitance and the shape of the anode bars, which were considered to be critical to voltage-tunable operation. Thus, in addition to the usual consideration of current, voltage, frequency, magnetic field and in-

[4] Radio Research Laboratory Staff, Harvard University, "Very High Frequency Techniques," vol. II, McGraw-Hill Book Co., Inc., New York, N. Y.; 1947.

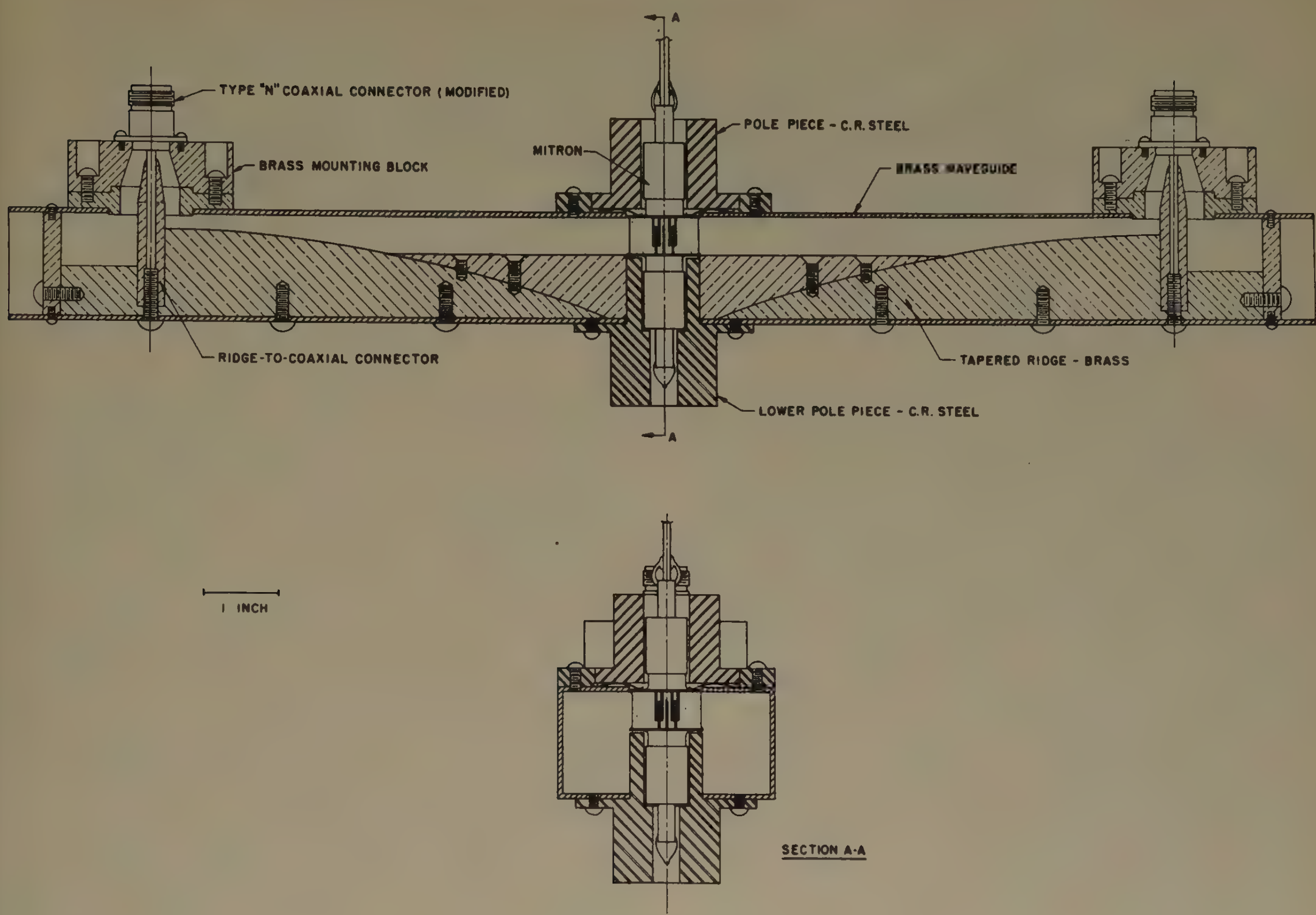

Fig. 1—Dual output cavity for mitron.

teraction space dimensions, special attention was given to these two parameters.

Hull[5] has suggested that the operation of a typical magnetron can be improved by redesigning the interaction space to reduce the higher order space (or Hartree) harmonics of the rf field. On the basis of these considerations, and the fact that round anode bars give less anode-to-anode capacitance, it was decided to use round anode bars in the design of the mitron.

Listed below are the values of the parameters which resulted from the design of the interaction space.

$f = 3{,}000$ mc (chosen to suit available test equipment)
$P_i = 50$ watts
$I_p = 0.050$ ampere (required to satisfy the maximum power demand at an anode voltage of 1,000 volts)
$N = 12$ (number of anode bars, based on existing magnetrons)
$r_c/r_a = 0.75$
$d_a = 0.030$ inch (diameter of anode bars)
$r_a = 0.205$ inch (determined by size of molybdenum rod available, and anode radius of existing magnetrons)
$L_c = 0.150$ inch (length of emitting surface on cathode required to satisfy maximum power requirements with a safety factor of two)
$l_s = 0.290$ inch (overlap length of anode bars; chosen to allow for length of emitting portion of cathode and end hats)
$E_0 = 250$ volts
$B_0 = 550$ gauss
$C_p = 0.9$ $\mu\mu f$ (anode-to-anode capacitance, no cathode present).

The Magnetic Circuit: In the design of the magnetic circuit there were two factors which were considered to be important to the successful operation of the tube: (1) that the magnetic field be as nearly uniform as possible throughout the interaction space, and (2) that relatively high fields, in the order of 3,000 gauss, be obtained without unreasonable requirements for an electromagnet. Fig. 2 (next page) shows the mitron with an oxide-coated cathode in place. As indicated, the iron pole pieces (part 1) serve as anode supports. The molybdenum anode rods are brazed onto the pole pieces. Parts 2 and 3 are made of Kovar to facilitate making the glass seals as indicated.

[5] J. F. Hull, "Analysis of a Magnetron Interaction Space for Elimination of Hartree Harmonics," Memo. Rep., SCCSCL-RTB2, Proj. 322A, Evans Signal Lab.; January, 1951.

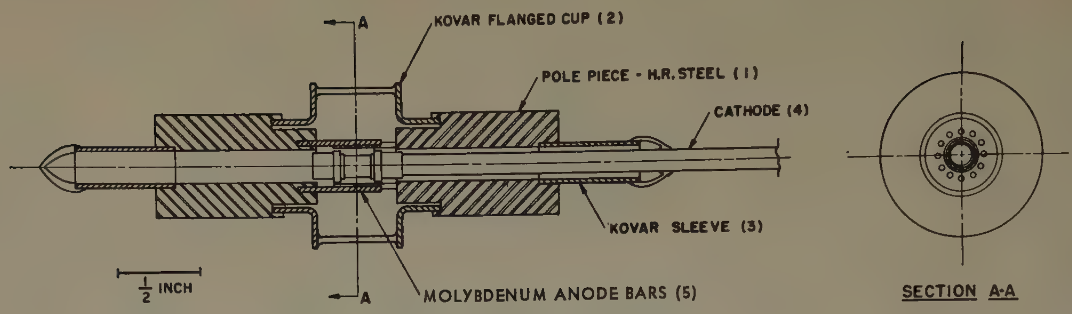

Fig. 2—Mitron.

Performance of the Mitron

Several types of cathodes have been used in the mitron; namely, oxide-coated, thoriated tungsten, tungsten, and a button cathode. Results of this investigation indicate, as expected, that the pure tungsten cathode is most satisfactory for use in the mitron. The first tube constructed used an oxide-coated cathode. This tube could not be operated cw because backheating prevented control of the cathode temperature required for coherent operation of the tube. The thoriated tungsten cathode made possible the cw operation of the tubes but was more sensitive to backheating than the pure tungsten cathode. The tube using a button cathode could only be operated under pulsed conditions because of the backheating of the cathode. For coherent operation of the tube, it was necessary to operate the cathode under temperature-limited conditions.

Typical data from a mitron which used a tungsten cathode are presented. This tube was tested under dynamic operating conditions. Results are shown in Figs. 3 and 4 (opposite). The conditions of operation are indicated on the figures. A 60-cycle voltage was superimposed upon the dc anode potential, giving an anode-voltage swing of 900 volts. Filament power was supplied from a battery to eliminate the frequency modulation caused by the variation of the magnetic field which is produced by an ac heater current. Fig. 3 shows the variation of the frequency with the anode voltage.

In addition to the voltage-tunable mode indicated in Fig. 3, there were two other modes which were very weak, but still detectable over portions of the voltage range. One of these modes tuned from 2,845 mc at 2,100 v to 3,170 mc at 2,300 v; the second tuned from 4,080 mc at 2,050 v to 4,370 mc at 2,200 v. In the voltage range 2,350 to 2,850 only the main mode was present. Under certain conditions of operation, there are discontinuities in the operation of the main mode. These discontinuities in the operation are detectable in either the power output or anode current curves. Frequency deviations of 400 to 600 mc can be obtained with no difficulty in maintaining continuous operation; for frequency deviations of the order of 800 to 1,000 mc, the adjustment of the cathode temperature becomes more critical.

With the modulation voltage reduced to 450 v peak-to-peak, the frequency of operation could be shifted as much as 50 mc by varying the filament heater current and still maintain coherent operation over the voltage range. An increase in cathode temperature decreases the frequency of oscillation. The tube seemed to operate best at, or near, the lowest cathode temperature.

For cw operating conditions, it was possible to vary the filament current from 8.9 to 9.3 amperes and still maintain a coherent signal output. The power output varied from 52 mw at an anode current of 5 ma to 350 mw at an anode current of 18.75 ma. The corresponding frequency variation was from 2,450 mc to 2,320 mc.

Fig. 4 shows the variation in power output and anode current with the anode voltage. The frequency range in Fig. 4 is approximately 900 mc (as shown in Fig. 3) and the peak power is of the order of 550 mw. Fig. 4 shows the variation of anode current with anode voltage, the average current being approximately 12 ma. It should be noted that there are no discontinuities in the operation and it was determined that the tube was oscillating coherently throughout the range. The decrease in the power output in the right (high frequency) side of Fig. 4(a) is due to the loading resulting from the cathode circuit at that frequency.

This tube was operated with battery supplies for the filament and anode, and a regulated power supply for the electromagnet. Under these conditions, the output signal (as observed on a spectrum analyzer) was stable; no jitter was present in cw operation. The output could be tuned over several hundred mc without readjusting the cathode temperature.

Satisfactory operation was obtained with anode voltages as low as 800 v and frequencies in the order of 2,000 mc. Under these conditions the output power was reduced, but the signal was stable and relatively free of noise.

A modification was made in the cathode helix so that the resonance between the helix and the coaxial filament lead was outside the frequency range of tube operation. Typical data from tubes using this cathode are shown in Fig. 5 (facing page). There, the frequency was varied from 2,190 mc to 3,330 mc, by applying a sine wave modulating voltage with an amplitude of 530 v peak-to-peak. The power output at the center of the range was 185 mw. It was determined that the tube was oscillating coherently throughout the range; there were no discontinuities in operation, and only one mode of operation was detectable.

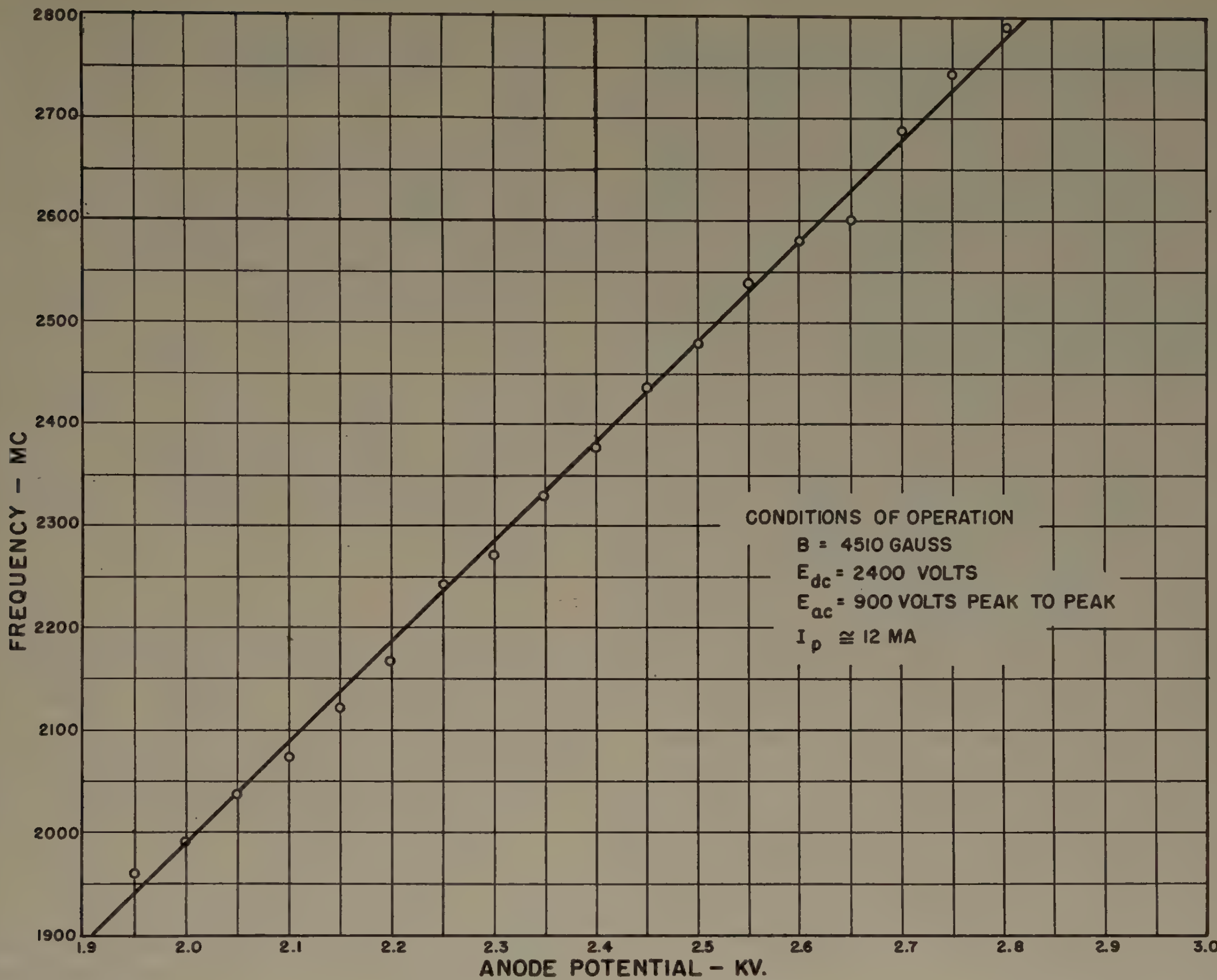

Fig. 3—Frequency vs anode potential.

An electronic controller has been used to maintain the dc anode current constant by controlling the filament power. This arrangement makes it possible to operate the mitron cw in the frequency range indicated in Fig. 5 without requiring manual readjustment of parameters other than anode potential. The tubes have been operated with permanent magnets, dc filament supply, and voltage regulated anode supply; under these conditions the output signal is stable. The output is sensitive to mismatches in the rf circuit, in some cases causing spurious oscillations and discontinuities in operation to occur. With a well-matched circuit, a single mode of operation is obtained with no discontinuities.

The mitron has been used as the local oscillator of a microwave receiver. The noise figure of the receiver (which consisted of a wide-band crystal mixer, a 30 mc IF amplifier with a 3 db bandwidth of 2 mc, and a de-

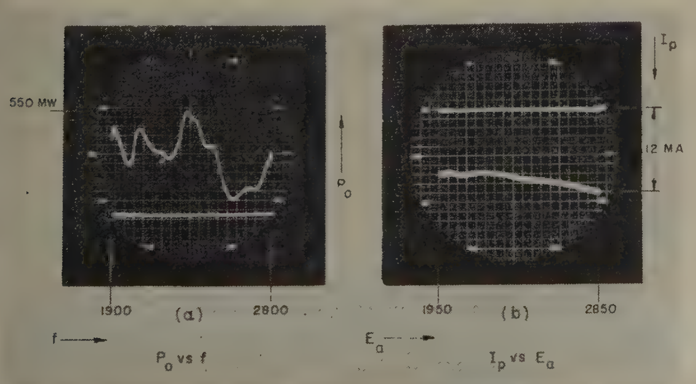

Fig. 4—Power and current characteristics.
B = 4,510 gauss E_{dc} = 2,400 volts
E_{ac} = 900 volts peak-to-peak I_p ≅ 12 ma.

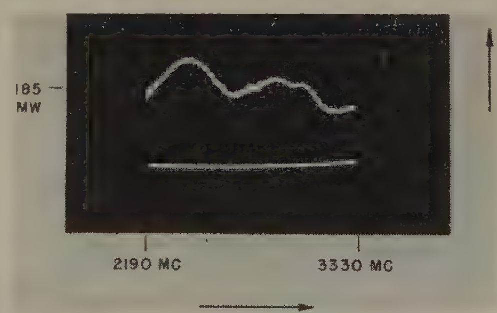

Fig. 5—Power vs frequency characteristics.
B = 2,200 gauss E_{ac} = 530 volts peak to peak
E_{dc} = 1,425 volts I_p ≅ 13.6 ma.

tector) was measured, using first the mitron and then a reflex klystron as the local oscillator. When a klystron was used as the local oscillator a noise figure of 8.5 db was obtained. When the mitron was used as the local oscillator, noise figures of 10.5 to 12.5 db were obtained. These variations in the noise figure have been attributed to variations in the anode current. The noise figure obtained with the mitron was sensitive to the anode current; a change in the anode current of 0.3 ma from the optimum caused an increase of 6 db in the noise figure. Measurements have been made in the frequency range 2,600 to 3,000 mc. The noise figure obtained with the mitron does not change much with frequency.

Effects of Circuit Design on Performance

The power output of the mitron has been shown to be very closely related to the total shunt impedance of the rf circuit, the power being a minimum when the shunt impedance is a minimum. With a wide-band nonresonant circuit, the peak power output was in the order of 500 mw; however, when the shunt impedance of the rf circuit was increased by shorting one end of the cavity, power output of 4 w was obtained. Under these conditions the frequency of the output signal could be varied by as much as 200 mc by varying the anode voltage.

Fig. 6 shows an approximate equivalent circuit for the mitron and its associated cavity. Fig. 6(a) is a first approximation in which the various elements are defined as follows: I_g is a current generator representing the rf current induced in the circuit by the space charge. θ is the phase angle between the rf voltage and rf current, the relation being

$$|Y_T|\,\underline{|\theta} = \frac{I_g\,\underline{|\theta}}{E_{rf}},$$

where I_g and E_{rf} are rms values, θ being defined so that it is negative for a lagging current.

C_p is the total capacitance between anode sets, equal to approximately 1 $\mu\mu f$.

L_p is the equivalent inductance of the anode bars between the interaction space and the anode supports, equal to approximately 130 $\mu\mu h$.

X_c is the impedance of the coaxial line formed by the cathode support and the pole piece looking out from the cathode; its value given approximately by

$$X_c = -jZ_{0_1}\cot\frac{2\pi l}{\lambda}, \tag{1}$$

where

$Z_{0_1} = 28.5$ ohms (characteristic impedance of the line)
l = length of the line.
λ = wavelength.

This assumes losses in the line, and losses due to radiation from extension of the center conductor, to be negligible.

C_1 and C_2 are the capacitances between end hats and respective anode sets.

G_t is the equivalent conductance representing the losses in the waveguide portion of the structure.

Z_1 is the rf impedance of cathode surface between end hats.

Z_2 is the rf impedance of the coaxial line formed by the filament leads.

T_1 and T_2 represent the waveguide-to-coaxial junctions, assumed to be ideal transformers.

Zl_1 and Zl_2 are the external loads connected to the coaxial output terminals.

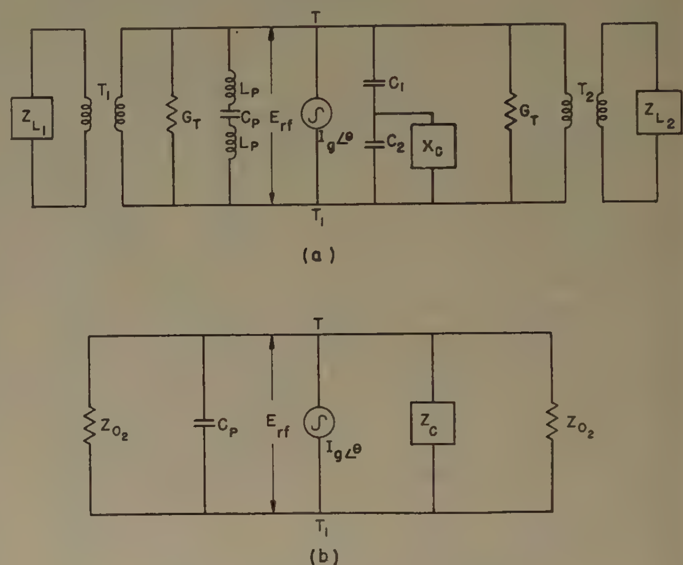

Fig. 6—Equivalent circuit of mitron.

In Fig. 6(a) (above) it has been assumed that $Z_1 = 0$; in Fig. 6(b), a helical cathode is assumed which gives a finite value to Z_1. Fig. 6(b) represents a second approximation in which it has been assumed that the coaxial output terminals are terminated in their characteristic impedances and that the losses in the waveguide are negligible compared to the load. Since the reactance due to $2L_p$ is never more than 10 per cent of the reactance due to C_p, X_{L_p} is neglected. Z_{0_2} is the characteristic impedance of the ridge waveguide, assumed to be a pure resistance given by

$$Z_{0_2} = Z_\infty \frac{1}{\sqrt{1 + \left(\frac{f_c'}{f}\right)^2}}. \tag{2}$$

Let

$$Z_c = X_1 + X_2 + Z_3,$$

where

$$X_1 = \frac{1}{j\omega C_1}$$

$$X_2 = \frac{\frac{1}{j\omega C_2} Z_c}{Z_c + \frac{1}{j\omega C_2}}$$

$$Z_2 = -jZ_{0_3} \cot \frac{2\pi l}{\lambda}$$

$$Z_3 = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

$Z_{0_3} = 39.2$ ohms (characteristic impedance of coaxial filament line).

The admittance of the circuit shown in Fig. 6 may therefore be expressed as

$$Y_T' = \frac{2}{Z_{0_3}} + j\omega C_p + \frac{1}{Z_c}. \quad (3)$$

Since Z_c is a complex impedance, it is this term in (3) which causes the greatest variation in total impedance with frequency.

Effect of Cathode on Performance

Experience has shown that an oxide-coated cathode is unsatisfactory for coherent voltage-tunable operation of the mitron. It has been possible to obtain coherent voltage-tunable operation using oxide cathodes. However, the mode of operation was very sensitive to the cathode temperature, and the operation of the tube was unstable due to the backheating of the cathode.

A pure tungsten cathode which operates at approximately 2,200 degrees Centigrade was found to be most suitable for voltage-tunable operation for obtaining a coherent signal. The operation of the mitron is sensitive to the temperature of the cathode, but the filament input power can be adjusted to give stable, coherent operation.

The improvement in voltage-tunable operation of the mitron with the use of a pure tungsten cathode is attributed primarily to the fact that more filament power is required for the pure tungsten cathode than for either the thoriated tungsten or oxide cathode. The emission from both oxide and thoriated tungsten cathodes is unstable and tends to cause the anode current to "run away." It is impossible to determine from presently available data the exact operating conditions at the cathode for voltage-tunable operation. It has been determined experimentally that the mode of operation is sensitive to the cathode temperature and that the mitron anode current varies with the temperature of the cathode.

Requirement for Temperature-Limited Operation of the Cathode

In the mitron there is a range of anode current for which a coherent signal is produced; for values of anode current above and below this range noncoherent signals are produced. The sensitivity of the mitron operation to cathode temperature may be attributed primarily to the low impedance which the anode and cavity structure presents to the electron stream.

Experimental data, obtained from Model 9 (bar-and-vane structure[3]) and mitrons, indicate that one of the most important parameters in obtaining voltage-tunable operation is the impedance presented to the electron stream by the anode and cavity structure. Coherent operation was obtained in the Model 9 magnetron with anode currents in the order of 100 μa. The operation was very sensitive to cathode temperature and coherent operation was obtained over a very narrow range of anode current. Coherent operation of a mitron has been obtained with anode currents in the order of 5 to 20 ma. The interaction space in the two tubes is quite similar. The most significant differences are in the anode-to-anode capacitance and in cavity impedance. The anode-to-anode capacitance in the Model 9 is approximately 2 $\mu\mu f$ and in the mitron approximately 1 $\mu\mu f$; the cavity impedance in the Model 9 is 25 ohms and in the mitron it is 80 ohms.

It is believed that voltage-tunable operation of a magnetron can be obtained under space-charge limited operation of the cathode if the impedance of the cavity and anode structure can be made sufficiently large.

Frequency vs Anode Voltage Characteristics for Voltage-Tunable Magnetrons

In order to compare the voltage-tuning characteristics of all tubes, an equation describing the frequency-anode voltage characteristic for a voltage-tunable magnetron has been derived. This equation is arranged in nondimensional form so that all experimental data can be compared with the theoretical expression by using a graphical presentation. This derivation is based on the equations for the high-Q magnetron; however, the results obtained should approximately describe the behavior of the low-Q voltage-tunable magnetron.

By combining the following well-known expressions:

$$E_H = B\frac{\omega_n}{2} r_a^2\left(1 - \frac{r_c^2}{r_a^2}\right) - 1/2\left(\frac{m}{e}\right) r_a^2\omega_n^2 \quad (4)$$

(Hartree equation)

$$E_u = \frac{B^2 e}{8m} r^2\left(1 - \frac{r_c^2}{r_a^2}\right)^2 \text{ (Hull cut-off potential)} \quad (5)$$

$$B_0 = \frac{2\dfrac{m}{e}\omega_n}{\left(1 - \dfrac{r_c^2}{r_a^2}\right)} \quad (6)$$

$$E_0 = \frac{m}{2e} r_a^2\omega_n^2 \text{ (synchronous energy equation)} \quad (7)$$

$$\frac{E_H}{E_0} = \frac{2B}{B_0} - 1 \text{ (normalized Hartree equation)}, \quad (8)$$

we obtain[6]

[6] J. A. Boyd, "A Voltage-Tunable Magnetron for Operation in the Frequency Range 1500 to 3000 Megacycles," Univ. of Michigan, Electron Tube Lab. Tech. Rep. No. 15; November, 1953.

$$\frac{f}{f_0} = \left[1 - \sqrt{1 - \frac{E_H}{E_u}}\right], \tag{9}$$

where

E_H = threshold potential
E_u = Hull cut-off potential
f_0 = is the frequency of the current induced in the anodes by an electron at the anode with a tangential energy of E_u.

$$f_0 = \left[\frac{E_u e}{m 2\pi^2}\right]^{1/2} \frac{n}{r_a}. \tag{10}$$

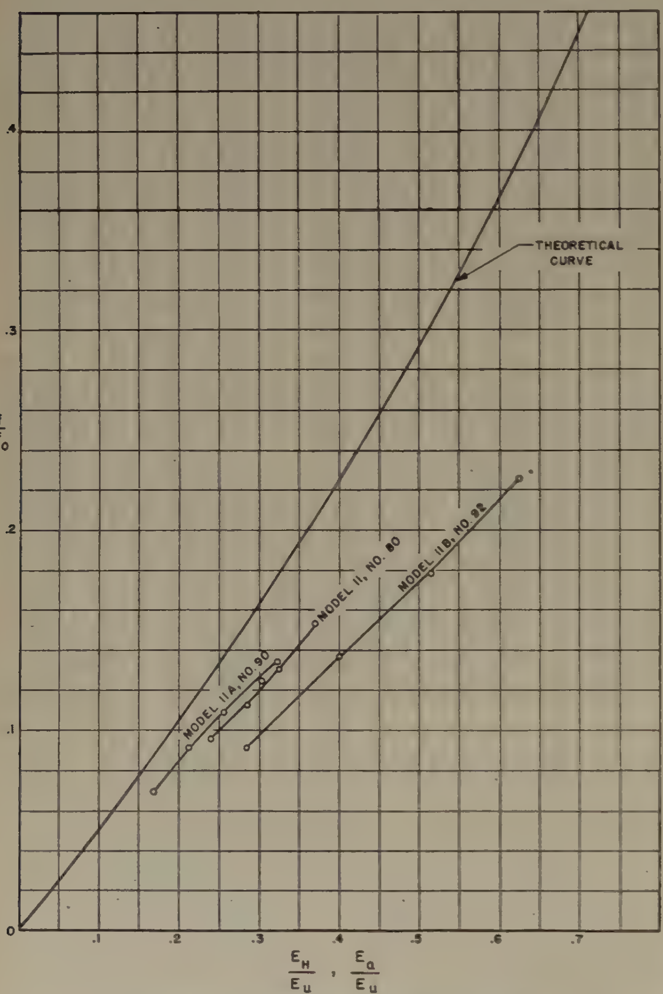

Fig. 7—Typical voltage tuning characteristics for Models 11, 11A, and 22B. π-Mode of operation assumed $n = N/2 = 6$.

Fig. 7 shows the theoretical tuning curve plotted for comparison with the experimental curves for three forms of the tube. In plotting the experimental data, f is the operating frequency and E_a is the anode potential; E_u and f_0 were determined from the operating conditions assuming π-mode of operation. It should be noted that all forms of the tube had essentially linear-tuning characteristics and all tubes operated at higher anode potentials than indicated by the Hartree equation, when π-mode of operation is assumed. It was assumed that the mitron would operate in the π-mode. Rotating-probe measurements on this tube and its associated cavity structure show that with an rf voltage applied to one of the output terminals of the cavity, a π-mode voltage distribution does exist on the anode bars.

It can be shown that if one assumes the presence of radial energy of the electrons that threshold potentials above the Hartree potential will be obtained.[6] There is experimental evidence that electrons in the mitron do possess considerable radial energy; therefore, it is concluded that the high operating potentials are at least in part due to the fact that the electrons arrive at the anode with considerable radial energy.

Conclusions

A voltage-tunable magnetron has been developed with operating characteristics that make the tube useful as a source of microwave power. It has been shown that the tube is satisfactory as the local oscillator of a spectrum analyzer and as an rf power source for most rf measurement applications. Noise figure measurements indicate that by using a balanced mixer it may be feasible to use the mitron as the local oscillator in microwave receivers.

A pure tungsten cathode was found to give better voltage-tunable operation than either an oxide-coated or a thoriated-tungsten cathode. The cathode support and the filament leads were found to cause variations in the power output of the tube.

With the present circuit arrangement, the anode-to-anode capacitance appears to be the limiting factor in obtaining higher power from the tubes. A reduction in the value of this capacitance is necessary if significantly higher power is to be obtained. Some improvement in efficiency and power output may be obtained by increasing the external circuit impedance although the improvement to be obtained in this way is limited.

The operating anode voltage was found to be higher than predicted by the Hartree equation, assuming π-mode of operation. This increase in operating voltage is attributed in part to the electrons reaching the anode with considerable radially-directed velocity. It is proposed that the sensitivity of voltage-tunable operation to cathode temperature is due primarily to the low impedance which is presented to the electrons by the anode and cavity structure.

Acknowledgment

The author wishes to acknowledge the contribution to this work made by other members of the Department of Electrical Engineering staff. Profs. W. G. Dow, G. Hok, H. W. Welch, Jr., J. S. Needle, and Mr. J. R. Black have contributed much by their helpful criticisms and discussions. The tubes and test equipment were constructed in the Electron Tube Laboratory, under the supervision of J. R. Black.

Resonant Behavior of Electron Beams in Periodically Focused Tubes for Transverse Signal Fields*

R. ADLER†, FELLOW, IRE, O. M. KROMHOUT†, ASSOCIATE, IRE, AND P. A. CLAVIER†, MEMBER, IRE

Summary—A ribbon beam which is focused by a series of bidimensional electrostatic lenses exhibits a well-defined transverse resonance within a certain range of focusing conditions. The frequency of the resonance is derived for the case of a simple space-periodic focusing field; it is also shown that the focusing field determines the current density in the beam (and thus, in practice, the maximum usable beam current). An experiment is described which demonstrates by means of a transverse signal field the existence of the transverse electron resonance.

SPACE-PERIODIC fields symmetrical with respect to a center plane x, z are of interest in tubes which use ribbon or sheet beams. A simple solution exists for the motion of an electron, traveling initially with the velocity $u_0 = dz/dt$ at the distance y_0 from the center plane, through a stationary field of the form

$$V = V_0(1 + \epsilon \sin \nu z \cosh \nu y), \tag{1}$$

under these two restrictions: (weak field)

$$\epsilon \ll 1 \tag{2}$$

and (electron close to axis)

$$y \ll \frac{1}{\nu}. \tag{3}$$

The derivation proceeds as follows: If ϵ were zero, the electron would continue along a straight line $y = y_0$. If ϵ is finite but very small, two first-order effects appear: a periodic variation of the velocity component $\partial z/\partial t$ about the average u_0, and a periodic variation of y (ripple) about the average y_0. Each of these perturbations cooperates with the periodic field to produce a second-order effect—a force component which, when averaged over one spatial period, is different from zero. These two average force components can be determined separately and added.

The following equations are needed in the derivation: From (3) follows

$$\cosh \nu y \sim 1 + \tfrac{1}{2}(\nu y)^2. \tag{4}$$

The dc potential V_0 is related to the average velocity u_0 as follows:

$$2\frac{e}{m} V_0 = u_0^2. \tag{5}$$

Contribution from variation of the velocity component $u = \partial z/\partial t$ is considered first. The instantaneous velocity is

$$u = u_0\left(1 + \frac{\epsilon}{2} \sin \nu z\right). \tag{6}$$

Hence

$$\frac{dt}{dz} = \frac{1}{u} = \frac{1}{u_0}\left(1 - \frac{\epsilon}{2} \sin \nu z\right). \tag{7}$$

Transverse field at distance y_0 from axis:

$$\frac{\partial V}{\partial y} = V_0 \epsilon \sin \nu z \cdot \nu^2 y_0. \tag{8}$$

Instantaneous transverse force:

$$f = e\frac{\partial V}{\partial y} = eV_0\epsilon \sin \nu z \cdot \nu^2 y_0. \tag{9}$$

Transverse impulse per unit length of travel:

$$f\frac{dt}{dz} = eV_0\epsilon \sin \nu z \cdot \nu^2 y_0 \cdot \frac{1}{u_0}\left(1 - \frac{\epsilon}{2} \sin \nu z\right). \tag{10}$$

Time average of force over one full period of $\sin \nu z$

$$= \text{time average of } eV_0\epsilon \sin \nu z \cdot \nu^2 y_0 \left(1 - \frac{\epsilon}{2} \sin \nu z\right)$$

$$= -\tfrac{1}{4} eV_0 \epsilon^2 \nu^2 y_0. \tag{11}$$

Contribution from periodic variation of y (ripple) is considered next. The transverse force from (9),

$$f = eV_0\epsilon \sin \nu z \cdot \nu^2 y_0,$$

can be expressed as a function of time:

$$f = eV_0\epsilon \sin u_0 \nu t \cdot \nu^2 y_0. \tag{12}$$

The expression y_0 now designates the distance from the axis averaged over one ripple in the trajectory caused by the periodic force. The ripple Δy is obtained by dividing (12) by the electron mass m and integrating twice with respect to time:

$$\Delta y = -\frac{1}{u_0^2 \nu^2} \frac{e}{m} V_0 \epsilon \sin u_0 \nu t \cdot \nu^2 y_0, \tag{13}$$

or, substituting from (5),

$$\frac{e}{m u_0^2} = \frac{1}{2V_0}:$$

$$\Delta y = -\tfrac{1}{2}\epsilon \cdot \sin u_0 \nu t \cdot y_0; \tag{14}$$

or in terms of z:

* Original manuscript received by the IRE, June 23, 1954; revised manuscript received October 29, 1954.
† Zenith Radio Corp., Chicago, Ill.

$$\Delta y = -\frac{\epsilon}{2}\cdot \sin \nu z\cdot y_0. \tag{15}$$

Thus the trajectory perturbed by ripple will be

$$y = y_0 + \Delta y = y_0\left(1 - \frac{\epsilon}{2}\sin \nu z\right). \tag{16}$$

Transverse force (9) is proportional to y. Taking this into account to get the second-order term, one finds

$$e\frac{\partial V}{\partial y} = eV_0\epsilon \sin \nu z\cdot \nu^2 y = eV_0\epsilon \sin \nu z\cdot \nu^2 y_0 \cdot\left(1 - \frac{\epsilon}{2}\sin \nu z\right). \tag{17}$$

This has the time average (taken over one period of νz)

$$-\tfrac{1}{4}eV_0\epsilon^2\nu^2 y_0. \tag{18}$$

The two contributions are equal and add up to $-\frac{1}{2}eV_0\epsilon^2\nu^2 y_0$. This has the form of an elastic restoring force; it could be ascribed to a fictitious electric field E_e (in the following called the elastic field) of the form

$$E_e = -Ky_0 = -\tfrac{1}{2}V_0\epsilon^2\nu^2 y_0. \tag{19}$$

The gross behavior of the electron, disregarding the small ripples and velocity variations which correspond to $\sin \nu z$, is determined by this elastic field. In such a field an electron performs transverse oscillations of the form $y = y_0\cdot\cos \omega_e t$ where ω_e, the angular frequency of transverse electron resonance, is

$$\omega_e = \sqrt{K\frac{e}{m}} = \sqrt{\frac{1}{2}\frac{e}{m}}V_0\cdot\epsilon\nu, \tag{20}$$

and by substituting (5),

$$\omega_e = \frac{\epsilon}{2}u_0\nu. \tag{21}$$

Dividing by u_0, we find $\omega_e/u_0 = (\epsilon/2)\nu$. The length of one wave of the trajectory caused by the transverse oscillations is $\lambda_e = 2\pi u_0/\omega_e$; the spatial period of the field is $s = 2\pi/\nu$; thus we obtain

$$\lambda_e = s\cdot\frac{2}{\epsilon}\cdot \tag{22}$$

Since $\epsilon \ll 1$, the concept of many spatial periods per wavelength is justified.

So far, this derivation has been concerned with a single electron. Let us now assume a sheet beam of charge density ρ, centered about $y = 0$ and thin enough to meet (3). The space charge produces a field

$$\frac{\partial V}{\partial y} = \rho y. \tag{23}$$

If the beam is stable, the elastic field must balance the space charge field:

$$\tfrac{1}{2}V_0\epsilon^2\nu^2 = \rho. \tag{24}$$

Also, since $\omega_e^2 = \frac{1}{2}(e/m)V_0\epsilon^2\epsilon^2$ [from (20)], we can write

$$\omega_e^2 = \frac{e}{m}\rho. \tag{25}$$

Thus the transverse resonance frequency ω_e equals the plasma frequency (as was pointed out by D. A. Dunn of Stanford University). The elastic field acts somewhat like a stationary positive plasma, in the center of which a beam of equal negative density is stable.

If the beam is displaced from the center plane by y_B, the space charge field becomes

$$\frac{\partial V}{\partial y} = \rho(y - y_B). \tag{26}$$

The elastic field E_e is still

$$E_e = -\tfrac{1}{2}V_0\epsilon^2\nu^2 y.$$

Then if (24) holds, there remains only a residual field

$$\frac{\partial V}{\partial y} = -\rho y_B = -\tfrac{1}{2}V_0\epsilon^2\nu^2 y_B. \tag{27}$$

This acts upon the entire beam as if all electrons were at y_B. Thus, the entire beam as a unit is capable of resonant transverse motion according to (20) and (22), just like a single electron.

It should be remembered that these conclusions apply only under the restrictions $\epsilon \ll 1$ (weak field) and $y \ll (1/\nu)$ (narrow strip along center plane). The first restriction excludes the stop bands observed in stronger fields; the second rules out some tubes, because of their construction. One might thus think that the derivations had no practical application.

We have found, however, that the transverse resonance at ω_e dominates that class of tubes designed to work with transverse signal fields. In such tubes it is vital that all parts of the beam move together under the influence of a signal field. To insure this, the displacement of the beam caused by a signal should produce equal increments of force in all parts of the beam; this calls for $\partial E_e/\partial y = \text{const}$ or $E_e = y\cdot\text{const}$ at any given cross section z. This condition is met whenever restriction (3) holds, i.e. all parts of the beam are close to the center plane.

In addition, it is required that each trajectory wave include enough spatial periods so that the specific location of these periods on any particular wave has negligible influence on its length; if a trajectory wave included only one or two spatial periods, the relative phase of wave and spatial period at the starting point would strongly influence the shape of each trajectory, so that the wavelength would vary irregularly in space and in some cases also in time. The required relationship is insured by restriction (2) in view of (22).

If either of the two restrictions is violated, the effect of a transverse signal field at one point is soon converted into turbulence instead of concerted motion of the entire beam. Thus the two restrictions are charac-

teristic of tubes which use transverse signal fields such as the one illustrated in Fig. 1.

We have built a number of such tubes based on the structure illustrated. A thin flat ribbon beam travels in a space bounded by closely spaced electrode pairs which are held alternately at high and low potentials.

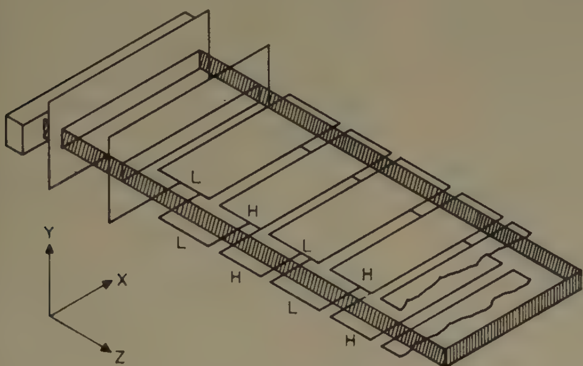

Fig. 1—Electrostatically focused ribbon-beam tube: beam from elongated cathode passes through collimating slots into space bounded by electrode pairs which are held at alternately high and low potentials.

Some of these tubes serve to study focusing; others are transverse-field traveling-wave tubes or related devices. We find consistently that the properties of these tubes are well described by the concept of a beam tied to the center plane by an elastic field and capable of resonant transverse oscillation.

One particularly convincing experiment is the following: In the structure illustrated, all plates on each side of the beam are interconnected by means of condensers so that it becomes possible to apply a homogeneous transverse signal field across the beam and deflect it from its normal path. After the beam has passed through this structure it is intercepted by a target assembly which, by means of an aperture of known width or by a fluorescent coating, provides information on how much the beam has been deflected. As the signal frequency is varied, a pronounced resonance is observed; its frequency is found to depend on the two dc voltages applied to the electrode pairs, but not on the beam current. In a typical case, the resonance occurs at 160 mc. A computation, using $V = V_0(1+\epsilon \sin \nu z)$ where $V_0(1 \pm \epsilon)$ are the two applied dc voltages (admittedly a crude approximation of the potential along the center plane of the structure shown), gives 142 mc. In this case, $\epsilon = .38$.

The maximum current which a tube can pass ranges from 30 per cent to over 50 per cent of that obtained by multiplying the density from (25) with the available cross section. (One would hardly expect 100 per cent, since thermal velocities are neglected in the derivation). We find the equations quite helpful in predicting the properties of such tubes.

The transverse electron resonance appears to permit some interesting practical applications on which we hope to report in the near future.

Correspondence

On Reciprocal Inductance*

Reference is made to the suggested term "inertiance" for reciprocal inductance, L^{-1} or Γ.[1] There are many models, making possible analogous mechanical and electrical systems. Electric networks have duals, and the possibility of more than one electric analog of a given mechanical system exists. One relationship has been chosen by Baghdady, and if this turns out to be the approved one, "inertiance" is an excellent term for L^{-1}. Using a different relationship this author arrived at the rather contradictory term "recimass."

It is important to note that the second formula given in the text referred to above

$$\frac{di}{dt} = \Gamma \frac{d\psi}{dt},$$

is a *fundamental* relation, not restricted to any particular model. When the table for transfer to a mechanical network is decided upon for one of the three quantities i, Γ, and ψ, the other two will adjust themselves accordingly. The formula cannot be used to prove that "inertiance" fits Γ better than it fits L. In an initial approach, would not most engineers associate "inertiance" with inductance rather than reciprocal inductance?

While "inertiance" supposedly solves half the nomenclature problem, it leaves the other half unsolved; the *unit* for L^{-1}. Capacitance has the unit "farad" and elastance the unit "daraf," both excellent. Inductance has the unit henry, but "inertiance" has no unit. This author suggested and used the unit *"yrneh"* at Cruft Laboratory in 1942, and now proposes this term for practical use.

There are many other symbols, terms, and abbreviations, which should be given proper identifications. A few of the ones the author has coined, more or less successfully, are as follows:

For immittance: ⊁, ⊁ (Z and Y combined)
For reciprocal immittance: ⊁, ⊁

For radius vector (which is not a conventional vector): radius spinner
For "a" in $\mathrm{H} = a + jb$: prime direction part Pr(H) or P(H). (The part "b" is widely referred to as the quadrature part, Qu(H) or Q(H).)
For amplification: amion, or ampion
For distributed amplifier: D—amplifier
For distributed power amplifier: DP—amplifier, or DPA

Other readers may independently have thought of these, or better ones, at an earlier date. While "amion" or "ampion" scarcely can be defended as abbreviations, one of them may survive, if considered as a new word. The reader's attention is called to the excellent suggestion by Professor R. Rüdenberg, Harvard University, of replacing the term "negative resistance" by "stimulance."[2]

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* Received by the IRE, December 27, 1954.
[1] E. J. Baghdady, "On reciprocal inductance," Proc. I.R.E., vol. 42, p. 1807; December, 1954.

[2] R. Rüdenberg, "Transient Performance of Electric Power Systems," McGraw-Hill Book Co., pp. 200, 467, New York; 1950.

The Mesh Counterpart of Shekel's Theorem*

In a recent paper Shekel[1] showed that the determinant of the admittance matrix of a network is independent of the choice of the reference node. The same statement is also made very often about the mesh determinant. The following example shows that the latter statement is not true.

Example: Choice 1 (see Fig. 1):

$$\Delta_1 = \begin{vmatrix} 3 & 1 & -1 \\ 1 & 3 & 1 \\ -1 & 1 & 3 \end{vmatrix} = 16.$$

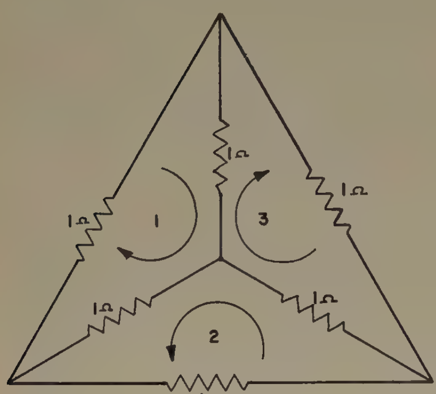

Fig. 1—Choice 1.

Choice 2 (see Fig. 2):

$$\Delta_2 = \begin{vmatrix} 4 & 0 & 0 \\ 0 & 4 & 0 \\ 0 & 0 & 4 \end{vmatrix} = 64.$$

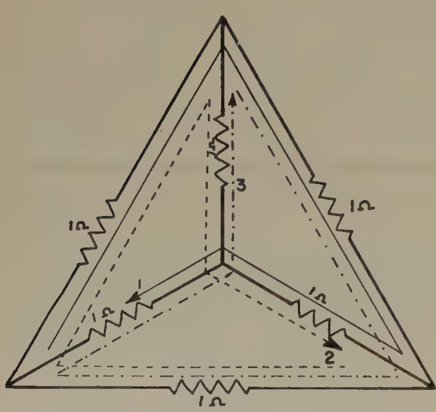

Fig. 2—Choice 2.

The following theorem gives the correct statement.

Theorem: If Δ_1 and Δ_2 are any two mesh determinants of the same network then $\Delta_2 = k\Delta_1$ where k is a real constant depending only on the two choices of circuits (and not on the impedances in the network).

Proof: The mesh impedance matrix Z_m is given by

$$Z_m = BZB',$$

where B is the circuit matrix of the network with circuits for rows and branches for columns (B is the transpose of the matrix Γ as defined by Cauer[2]), B' is the transpose of B, Z the branch impedance matrix. Let

$$Z_m = B_1ZB_1'$$

and

$$Z_{m_2} = B_2ZB_2'.$$

(Z is the branch impedance matrix and is thus independent of the choice of circuits.)

Since B_1 and B_2 form two bases for the same system of circuits there exists a non-singular matrix D such that[3]

$$B_2 = DB_1$$

B_1 has a rank equal to the number of rows ($b-n+1$ for a connected network). Hence B_1 contains a non-singular submatrix of maximum rank. Let the columns of B_1 be arranged such that this non-singular submatrix appears in the leading position. Let the columns of B_2 be arranged in the same order as the columns of B_1. Letting $B_{1_{11}}$ be the non-singular submatrix of B_1 we have

$$[B_{2_{11}}\ B_{2_{12}}] = D[B_{1_{11}}\ B_{1_{12}}]$$

from which

$$B_{2_{11}} = DB_{1_{11}}.$$

Since the product of nonsingular matrixes is nonsingular, $B_{2_{11}}$ is nonsingular. Also, since B_1 and B_2 contain only elements 1, -1, or 0, the determinants of $B_{1_{11}}$ and $B_{2_{11}}$ are both real constants. Hence using the notation det. A for the determinant of matrix A,

$$\text{det. } D = \text{det. } B_{2_{11}}/\text{det. } B_{1_{11}}$$

is also a real constant. Let this be r. Then

$$\Delta_2 = \text{det. } B_2ZB_2' = \text{det. } DB_1ZB_1'D'$$

or

$$\Delta_2 = \text{det. } D \times \text{det. } B_1ZB_1' \times \text{det. } D'$$

or

$$\Delta_2 = r \times \Delta_1 \times r$$

or

$$\Delta_2 = r^2\Delta_1 = k\Delta_1$$

where $k=r^2$ is a real constant. This completes the proof.

Since the mesh determinant is a rational function and the poles and zeros of a rational function are unaltered by a constant multiplier we also have this obvious corollary.

Corollary: The poles and zeros of the mesh determinant are independent of the choice of meshes.

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* Received by the IRE, July 19, 1954; revised manuscript received, September 13, 1954.

[1] J. Shekel, "Two theorems concerning change of voltage reference terminal," Proc. I.R.E., vol. 42, p. 1125; July, 1954.

[2] W. Cauer, "Theorie der linearen Wechselstromschaltungen," Akademische Verlagsgesellschaft, Leipzig, p. 33; 1941.

[3] G. Birkhoff and S. MacLane, "A Survey of Modern Algebra," The MacMillan Co., Theorem 1, p. 245, New York; 1953.

The Compensation Theorem*

The compensation theorem in network theory is generally very lightly treated in books as one of those theorems which have to be mentioned for the sake of completeness but which are obvious statements without much value. The cause for this deplorable fate can probably be attributed to the reasoning that, in order to apply the compensation theorem, a solution of the circuit under consideration should first be available and the theorem therefore, is merely of academic interest rather than usefulness.

This note aims to point out the usefulness of the compensation theorem and also supplies two proofs for the theorem.

One form of the compensation theorem can be stated as follows: The effect of a change ΔZ in the impedance of a network branch is equivalent to that of inserting a compensating potential source $(-I\Delta Z)$ in series with the changed branch, where I is the current which flowed in the branch before the change; the relative polarities of the voltage $(-I\Delta Z)$ are such that it represents a potential rise in the direction of I. The theorem expressed in this form is extremely useful in predicting the effect of changes in the value of an element on the performance of the whole circuit. Where only the effects of changes in different circuit elements are of importance, re-computation and subtraction by the original values are not the best approach. Computations in connection with dc amplifier stability are a good example, where, for small changes, the application of the compensation theorem makes possible the use of equivalent circuits for vacuum tubes. It is to be noted, however, that restriction of small changes is not required for validity of compensation theorem.

Proof 1: Consider the network in Fig. 1(a) where an impedance Z is connected between two network terminals A and B.

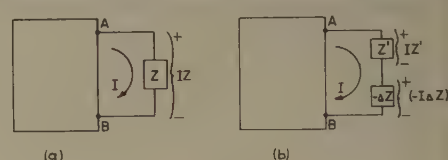

Fig. 1

The current I establishes a potential difference IZ between $A-B$ as shown. That Fig. 1(b) is equivalent to Fig. 1(a) is readily seen ($\Delta Z = Z'-Z$); what constitutes the network to the left of terminals $A-B$ is immaterial. The effect of the change of Z to Z' is then equivalent to that of cancelling the potential difference $(-I\Delta Z)$ by an equal and opposite potential source as shown in Fig. 2. This proves the theorem.

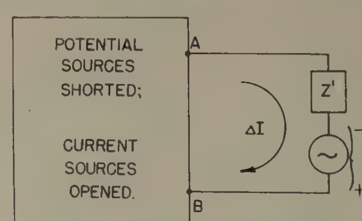

Fig. 2

Proof 2: Consider the general loop equation for the s-th loop of an n-loop network.

$$Z_{s1}I_1 + Z_{s2}I_2 + \cdots + Z_{ss}I_s + \cdots + Z_{sn}I_n = E_s, \quad (1)$$

where E_s is the algebraic sum of all applied potential rises in the direction of I_s in loop s. It is important to note that, if all loop currents are chosen in the same reference direction, e.g., clockwise, then Z_{ss} is a positive

* Original manuscript received by the IRE, October 25, 1954; revised manuscript received, December 13, 1954.

quantity and all other impedance coefficients for a flat network can be made negative by an appropriate choice of loops. Suppose the impedance in an outside branch (a branch in which only one loop current flows) in the s-th loop is increased by an amount ΔZ, then (1) will be changed to

$$Z_{s1}I_1' + Z_{s2}I_2' + \cdots + (Z_{ss} + \Delta Z)I_s' + Z_{sn}I_n' = E_s. \quad (2)$$

Subtracting (1) from (2), one has

$$Z_{s1}(I_1' - I_1) + Z_{s2}(I_2' - I_2) + \cdots + (Z_{ss} + \Delta Z)(I_s' - I_s) + \Delta Z I_s + \cdots + \cdots + Z_{sn}(I_n' - I_n) = 0. \quad (3)$$

Eq. (3) can be re-written as

$$Z_{s1}\Delta I_1 + Z_{s2}\Delta I_2 + \cdots + Z_{ss}'\Delta I_s + \cdots + Z_{sn}\Delta I_n = -\Delta Z I_s, \quad (4)$$

where $Z_{ss}' = Z_{ss} + \Delta Z$ is the new loop impedance for the s-th loop. It is readily seen that (4) leads directly to the compensation theorem as stated above. There is no difficulty in arriving at the same conclusion if the impedance change occurs in a common branch. Besides, a given branch in a network can always be made an outside branch by choosing the loops suitably.

It is also easy to prove (a) that the network determinant of impedance coefficients is changed from D_z to $(D_z + \Delta Z D_{ss})$ by the impedance change, where D_{ss} is the cofactor of D_z with the s-th row and s-th column omitted; and (b) that the change in the loop current in a k-th loop, ΔI_k, is

$$\Delta I_k = \frac{-\Delta Z I_s D_{sk}}{D_z + \Delta Z D_{ss}}. \quad (5)$$

In (5), all quantities except ΔZ on the right-hand side are in terms of known quantities before the change.

A parallel statement of the compensation theorem, which will be useful where the network is more conveniently analyzed on a junction basis, is: The effect of a change ΔY in the admittance of a network branch is equivalent to that of applying a compensating current source $(-E\Delta Y)$ in parallel with the changed branch, where E is the potential difference which appeared across the branch before the change; the direction of the current $(-E\Delta Y)$ is such that it flows toward the junction which has positive reference polarity. Eq. (5) would apply equally well to this case, if loop currents and impedances are changed to junction-pair voltages and admittances respectively.

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Some Antics in Semantics*

The only appropriate reply to M. S. Corrington[1] for his interesting exploration into mathematical semantics is

$$\frac{q}{\sqrt{\pi}}\int_0^\infty \frac{1}{\sqrt{x}}\epsilon^{-x/100}dx = 10q.$$

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* Received by the IRE, November 18, 1954.
[1] M. S. Corrington, "Adventures in Calculus," Proc. I.R.E., vol. 42, pp. 1703; November, 1954.

Logarithmic Amplifier Simplifications and Improvements*

The article by Chambers and Page[1] admirably describes one method of obtaining a logarithmic response. We, at Philco, have been over the same ground and find that simplifications are possible with no degradation of performance. For this reason, the following general discussion of the matter is offered. Much of this may be found in Patent #2,662,978, Logarithmic Transducer, D. E. Sunstein, and Patent #2,663,015, Object Location System Employing Logarithmic Transducer, D. E. Sunstein, both applied for June 11, 1946.

First, we have found that certain precautions are necessary if one is to obtain from each stage a video output vs IF input which saturates in the manner described in Fig. 3 of the article referred to.[1] That is, there is, in general, a tendency for the video detected output obtained from an IF amplifier to increase with increase of input signal level up to some particular maximum output value. Further increase in IF input signal may cause a reduction in video output. This effect has been particularly observed when plate detection alone is used to develop the video output signal from a given IF stage. When the sum of plate and screen current is used, however, to form the video output, this back down characteristic does not occur. Such a back-down characteristic, however, has also been found to occur when grid current is used for detection and when the grid load impedance is moderately high. In this instance, the back-down characteristic usually occurs as a result of overload in the preceding stage rather than of overload in the stage producing the video output being measured, but can be prevented by the use of small grid video load resistors.

Thus, we have found it practical to produce useful video detected signals either from the cathode current of the stage, the grid current of the stage, or the sum of the plate and screen current.

We have not found it necessary to cause the IF gain per stage to be reduced below that which would be obtained for a conventional IF amplifier. This is in contrast to the work of Chambers and Page who indicate about 4 to 5 db less gain per stage than obtainable with conventional practice. In the case of cathode detection, we have found it is not necessary to suffer a loss in gain simply by using a normal value of cathode resistance on the order of 100 ohms, rather than the relatively high value of 3,600 ohms used by Chambers and Page. In this instance, we can provide an adequate IF bypass across the cathode resistor without suffering video smearing. A further benefit as a result of the low cathode impedance which we have found practical will be realized later in that it plays a part in permitting the removal of the separate video isolation amplifier utilized by Chambers and Page.

Of the three useful methods of producing a video output from the overloaded IF stage, we have found that the use of plate plus screen current has not been plagued by any inability to compensate for end effects or difficulties with regard to IF feedback as experienced by Chambers and Page. However, plate plus screen detection has been found undesirable in some installations because of the presence of hum and assorted other interferences on the $B+$ line.

The use of cathode or grid detection removes the requirement for an extremely well filtered $B+$ line. It has been found, in practice, that grid detection usually cannot provide as high video frequency capability as cathode detection, without suffering some loss in IF gain. For this reason, the amplifiers we have made in the past several years have all used cathode detection, but with the grid returned directly to ground through a low impedance, and with a cathode resistance on the order of 100 ohms.

By providing a video output from each IF stage with an impedance of only 100 ohms or less, we have been able to linearly add the detector outputs in a delay line of impedance of perhaps 1,000 ohms simply by use of an isolating resistor of about 15,000 ohms between each video detected output and the corresponding tap on the delay line. This causes the signal on the delay line to be substantially lower than in the Chambers and Page receiver. Reduced gain can be made up at end of delay line by a single video amplifier rather than 5 or more video amplifiers feeding the several taps on the delay line.

Moreover, we have found that it has not been necessary to limit the choice of interstage coupling networks to synchronous single or multiple tuned systems. In fact, we have successfully used staggered tuning to achieve the usual advantages with regard to effective gain-bandwidth product. In order to use staggered tuning, the staggering should be so chosen that all the poles lie on a circle whose center is at the IF carrier frequency. In this way, the gain of each stage at carrier is identical, and this has been found sufficient to insure a clean transient response to radar pulse signals having any rate of rise to which the IF can respond. Moreover, it has been found that the use of such staggered tuning still provides a selectivity curve which is adequately symmetrical about carrier when measured at any given input signal strength, or even when measured with input signal varying an amount necessary to hold output constant.

By the use of these techniques, we have been able to obtain a 60 db logarithmic range utilizing four 6AK5's followed by a crystal video detector. In other applications, more than 90 db of dynamic range has been obtained utilizing a correspondingly increased number of detecting IF stages. These IF stages so constructed have, in fact, been in production for several years, and have utilized no more tubes than a standard IF having the same signal gain and bandwidth.

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* Received by the IRE, November 26, 1954.
[1] T. H. Chambers and I. H. Page, "The high-accuracy logarithmic receiver," Proc. I.R.E., vol. 42, pp. 1307–1314; August, 1954.

Rebuttal[2]

Mr. Sunstein's letter seems to indicate that he has missed the objective of the work leading to the development of the loga-

[2] Received by the IRE, December 9, 1954.

rithmic receivers described in our article. The parent problem required a logarithmic receiver having a characteristic accurate to within about $\pm 1/2$ db throughout its range, and having the extreme stability and reliability so essential in equipment for computer-type applications.

The authors are familiar with the work not only of the Philco Corporation but also with the work done by other American concerns and by the British.

As Mr. Sunstein points out, the use of the successive detection principle (which incidentally was in use at the Naval Research Laboratory as early as 1942) allows the use of almost any type of detector in the logarithmic receiver, provided that it is accompanied by satisfactory limiting or overload characteristics in the IF stages. However, extensive study of and experimentation with the various systems has indicated that only the nonamplified back-biased stage with its outputs summed through unilateral elements will yield the desired $\pm 1/2$ db accuracy without critical adjustments and despite tube changes, supply voltage variations and the other exigencies of service use.

Mr. Sunstein's use of low cathode resistances and (bilateral) resistances for performing the summing function is certainly entirely satisfactory for a lower accuracy receiver. It will, however, yield a characteristic which is sensitive to tube aging and changes and to other variations.

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Naval Research Laboratory
Washington 25, D. C.

Hysteresis in Klystron Oscillators*

Hysteresis in klystron oscillators falls into two categories. One is circuit hysteresis, the other is electronic hysteresis. Circuit hysteresis is due to such things as high-Q loads and long lines between klystron and load, and is well understood and adequately discussed in the literature.[1,2] The causes of electronic hysteresis are less unambiguous.

Perhaps the best known electronic cause of hysteresis is that of multiple transit electrons in reflex klystrons. This is partially treated both by Pierce and Shepherd[1] and by Hamilton.[2] A reasonable qualitative picture exists, but quantitative calculat'ons are formidable.

It is stated by Pierce and Shepherd[1] and by Hamilton[2] that another important cause of hysteresis is that due to phase shift of electronic admittance with rf voltage. This can be calculated from second-order bunching theory. A simple way of visualizing the way in which phase shift can produce hysteresis is by reference to the admittance diagram of a klystron oscillator (Figs. 1 and 2). It is seen that if the locus of the electronic admittance vector is a curved line as indicated, hysteresis with two breaks will occur on one side of the mode only. It is further observed that increasing the load conductance should move the hysteresis points toward the edge of the mode, and that for

* Received by the IRE, November 19, 1954.

[1] J. R. Pierce and W. G. Shepherd, "Reflex oscillators," *Bell Sys. Tech. Jour.*, vol. 26, pp. 460–681; July, 1947.

[2] D. R. Hamilton, J. K. Knipp, and J. B. H. Kuper, "Klystrons and Microwave Triodes," McGraw-Hill Book Co., New York, N. Y., pp. 384–440; 1948.

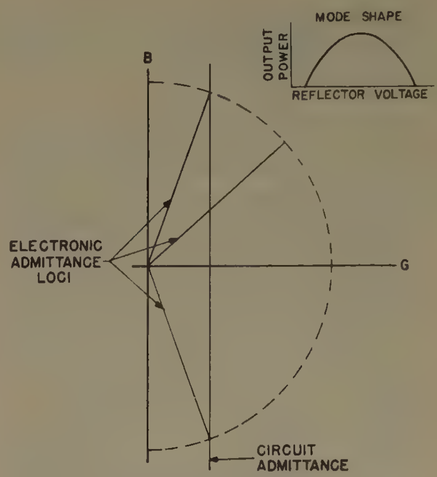

Fig. 1—Admittance diagram of klystron oscillator without phase shift.

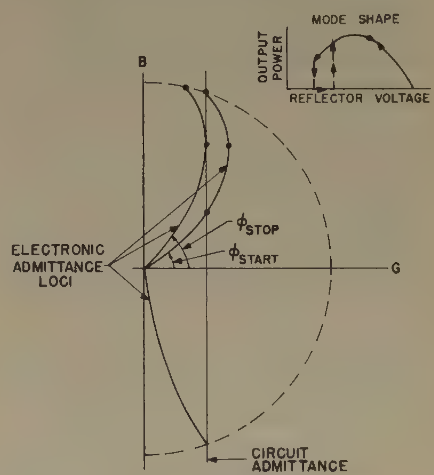

Fig. 2—Admittance diagram of klystron oscillator with phase shift.

large enough loads the hysteresis should disappear entirely.

It is found experimentally with certain reflex klystron oscillators, and with some two-cavity klystron oscillators, that a type of hysteresis occurs which cannot be explained in terms of either multiple transit electrons or simple phase shift. Examples of this kind of hysteresis are shown for a two-cavity oscillator in Fig. 3.

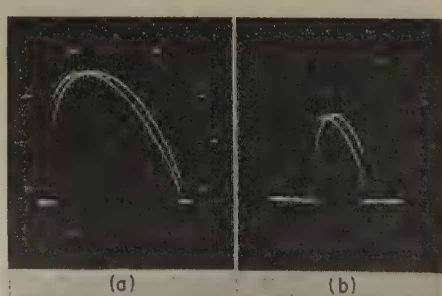

Fig. 3—Oscilloscope photographs of power vs beam voltage (sinusoidal sweep) for two-cavity klystron oscillator. (The two traces are displaced slightly in phase to show the hysteresis breaks.) (a) Moderate load. (b) Heavy load (large load conductance).

Dependence on load or circuit admittance is indicated. Multiple transit electron hysteresis cannot occur because there is no returned beam. While the observed hysteresis may be modified by phase shift, it is clear that phase shift alone is not adequate to produce the observed hysteresis. For one thing, the hysteresis occurs on both sides of the mode. In addition, the hysteresis points move toward mode center rather than away from it as the load conductance is increased.

The following explanation, supported additionally by other experiments, is proposed: It is usually supposed that, in the absence of multiple transit electrons, the electronic admittance decreases in magnitude monotonically with increasing rf voltage. Suppose, instead, that the electronic admittance first *increases* with rf voltage and then decreases, as indicated in Fig. 4. This is clearly a situation that will produce hysteresis on both sides of the mode, and will allow the break points to move toward mode center as the load conductance is increased.

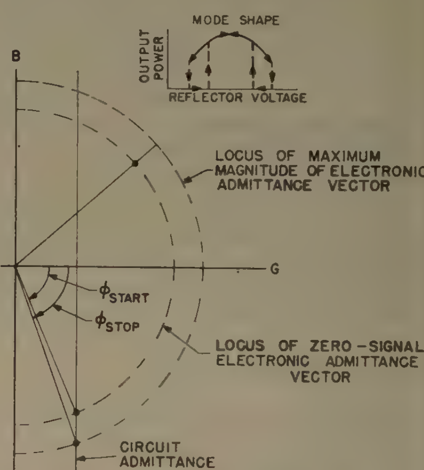

Fig. 4—Admittance diagram of klystron oscillator, with admittance first increasing with rf voltage and then decreasing.

There are several reasons why electronic admittance might increase with rf voltage before decreasing. One possibility is that rf causes a spreading of the beam, which may improve its coupling to the rf output gap (or increase the beam-coupling coefficient). Another possibility is that rf spreading decreases space-charge density and thus reduces certain deleterious space-charge effects in the beam, such as longitudinal debunching. In the case of the reflex klystron, reduction in space-charge density in the neighborhood of the turn-around plane in the reflector region may favorably affect the optics, thus increasing the electronic admittance. It might be expected that such space-charge effects would be more pronounced in klystrons having high density electron beams.

In two-cavity klystron oscillators, it is likely that both phase shift and growth of the electronic admittance vector occur simultaneously, thus causing a mixed type of hysteresis. In reflex klystrons, multiple transit electron effects often exist in addition, so that the resultant hysteresis is some combination of the three types. This obscures their separation and identification.

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Contributors

R. Adler (A'42–SM'47–F'51) was born in Vienna, Austria, in 1913. He received the Ph.D. degree in physics from the University of Vienna in 1937. He was assistant to a patent attorney in that city and later worked in the laboratories of Scientific Acoustics, Ltd., in London and Associated Research, Inc., in Chicago. In 1941 he joined the research group of Zenith Radio Corp. in Chicago, where he became associate director of research in 1952.

R. Adler

He is a member of the American Association for the Advancement of Science.

❖

H. G. Booker (SM'45–F'53) was born at Barking, Essex, England, in 1910. He was educated at Cambridge University, where he received the B.A. degree in 1933 and the Ph.D. degree in 1936. He was awarded the Smith's Prize in 1935 and became a Research Fellow of Christ's College in the same year.

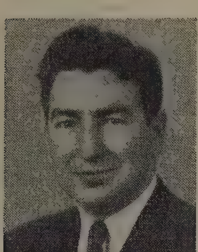
H. G. Booker

During the war he was in charge of theoretical research at the Telecommunications Research Establishment in England. After the war he returned to Cambridge University as a lecturer, and in 1948 became a professor of electrical engineering at Cornell University. He is now on leave of absence from Cornell and is at Cambridge University.

Dr. Booker is an associate member of the Institution of Electrical Engineers, London.

❖

J. A. Boyd (M'52) was born in Oscar, Kentucky, on March 25, 1921. He received the B.S.E.E. and M.S.E.E. degrees from the University of Kentucky in 1946 and 1949 respectively, and the Ph.D. degree from the University of Michigan in 1954. He was an instructor in electrical engineering in 1947–48 and an assistant professor in 1949 at the University of Kentucky.



J. A. Boyd

From 1949 to 1954 he has been employed as a member of the research staff and lecturer in the Department of Electrical Engineering at the University of Michigan. He is now an assistant professor in electrical engineering at the University of Michigan.

He is a member of Eta Kappa Nu, Tau Beta Pi and Sigma Xi.

❖

P. A. Clavier (M'48) was born in Paris, France, in 1922. He graduated from the Ecole Centrale des Arts et Manufactures de Paris in 1947. From 1947 to 1948 he was a research engineer at Laboratoire Central de Telecommunications in Paris where he worked on traveling-wave tubes.

P. A. Clavier

From 1948 to 1953 he was employed at the Physics Laboratory of Sylvania Electric, working on microwaves and applied mathematics. In 1952 he was appointed engineer in charge of the newly formed theoretical section. Since 1953 he has been with Zenith Radio Corp. in Chicago.

He is a member of the American Physical Society and the American Association for the Advancement of Science.

❖

For a photograph and biography of C. C. Cutler, see page 231 of the February, 1955 issue of the Proceedings of the I.R.E.

❖

J. T. deBettencourt (A'50) was born in Washington, D. C., on June 9, 1912. He received the B.E.E. degree in 1932 and the M.S. degree in physics in 1934 both from the Catholic University of America. He received the M.S. degree in 1937 and the Sc.D. degree in 1949 both in communication engineering at Harvard University.

deBettencourt

From October, 1941 to June, 1946, Dr. deBettencourt served with the Bureau of Ships, during the latter period as head of the Airborne Radar Design Section. After the war, he joined the Raytheon Manufacturing Co. as part-time engineer while finishing his Doctorate as research associate at Harvard. Until 1951 he was section head and section manager in the Engineering Division. In January, 1951, he joined the Research Laboratory of Electronics at M.I.T. as a staff member and is now with the Lincoln Laboratory.

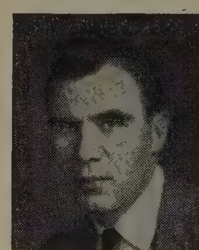
S. S. Guterman was born in Warsaw' Poland, in May, 1913. He received the equivalent of the American B.S. degree in electrical engineering from Polytechnic Institute of Grenoble, France. Later, he attended the Ecole Superieure of "Electricite" in Paris, where in 1935 he received a degree corresponding to the M.S. in telecommunication engineering.

S. S. Guterman

In 1951 he came to the United States, where, since May, 1952, he has been with the Raytheon Manufacturing Company as a group leader in the research and development section of the Computer Department.

❖

M. E. Hines (S'46–A'47–M'50) was born on November 30, 1918, in Bellingham, Wash. He attended the California Institute of Technology, where he received the B.S. degree in applied physics in 1940 and where, as a member of the Air Force Weather Service, he received the B.S. degree in meteorology in 1941. After World War II he returned to the Institute and received the M.S. degree in electrical engineering in 1946.



M. E. Hines

Mr. Hines has been employed since 1946, at the Bell Telephone Laboratories at Murray Hill, N. J., where he has been engaged in the development of traveling-wave tubes, microwave triodes and storage tubes.

❖

R. D. Kodis (S'46–A'51) was born on July 8, 1921, in Boston, Mass. He received the B.S. degree from Northeastern University in 1946.

R. D. Kodis

Mr. Kodis has been associated with the Harvard Computation Laboratory (1946–1947), Watertown Arsenal Laboratory (1948–1951), and Transducer Corp. (1951–1952). In 1952 he joined the Raytheon Manufacturing Company, where he is head of the research and development section of the Computer Department.

Contributors

O. M. Kromhout (A'54) was born in New Orleans, Louisiana, in 1925. She received a B.A. degree from Tulane University in 1945. After five years with the Department of Agriculture, she obtained an M.S. degree at the University of Illinois in 1951.

O. M. Kromhout

Mrs. Kromhout has been a research physicist with the Zenith Radio Corp. since 1952, where she is working on high frequency electron devices.

She is a member of Pi Mu Epsilon.

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For a photograph and biography of J. T. Mendel, see page 863 of the May, 1954 issue of the Proceedings of the I.R.E.

❖

Allen Nussbaum was born in Philadelphia, Pa., on August 22, 1919. He received the B.A. degree in chemistry in 1939 and the M.A. degree in physics in 1940, from the University of Pennsylvania. He entered service in 1941, serving as a radio engineer with the Signal Corps in Italy and with the Air Force in Germany from 1947 to 1949.

A. Nussbaum

Dr. Nussbaum reentered the University of Pennsylvania in 1950, where he received the Ph.D. degree in semiconductor physics in 1953. Since November, 1953, he has been a member of the semiconductor group at the Honeywell Research Center.

He is a member of the American Physical Society.

❖

Smil Ruhman was born on September 2, 1925, in Bessarabia, Roumania. He received the B.S. degree in electrical engineering from the Worcester Polytechnic Institute in 1949.

S. Ruhman

From 1949 to 1952, he served as research assistant at the Digital Computer Laboratory of the Moore School of Electrical Engineering, University of Pennsylvania, where he took part in the development of the MSAC, a large scale electronic digital computer. While at the University of Pennsylvania, he continued his studies in electrical engineering, completing the course requirements for the M.S. degree. In July, 1952, he joined Raytheon Manufacturing Company, Computer Department, where he has been engaged in research and development work in magnetics.

Mr. Ruhman is a member of Tau Beta Pi and Sigma Xi.

❖

J. A. Saloom (S'48–A'52) was born in Herrin, Illinois, in 1921. He attended the University of Illinois from 1942 to 1943, served in the U. S. Air Force from 1943 to 1946, and returned to the University in 1946 to continue his studies. He received the B.S., M.S. and Ph.D. degrees from the University of Illinois in 1948, 1949 and 1951 respectively. From 1949 to 1951 Dr. Saloom was a research associate at the University of Illinois, engaged in the study of gaseous discharges. In 1951 he joined the Bell Telephone Laboratories and has since been engaged in microwave tube development, with emphasis on electron beam studies.

J. A. Saloom

Dr. Saloom is a member of Pi Mu Epsilon, Eta Kippa Nu and Sigma Xi.

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For a photograph and biography of P. K. Tien, see page 82A of the September, 1954 issue of the Proceedings of the I.R.E.

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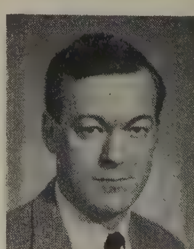
L. R. Walker was born in England, on August 6, 1914. He received the B.Sc. degree in 1935 and the Ph.D. degree in 1939 from McGill University. From 1939 to 1941 he was with the Radiation Laboratory at the University of California, and from 1941 to 1946 with the Radiation Laboratory at the Massachusetts Institute of Technology, where he worked on magnetrons. Since 1946 he has been a member of the technical staff of the Bell Telephone Laboratories, Murray Hill, N. J., engaged in work on microwave tubes and gyromagnetron propagation.



L. R Walker

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W. M. Webster (A'48–SM'54) was born in Warsaw, N. Y., in 1925. He studied physics at Rensselaer Polytechnic Institute, and at Union College. He received the B.S. degree in physics from Union College, and the Ph.D. degree from Princeton University.

W. M. Webster

In 1946 he joined the RCA Laboratories Division, Princeton, N. J., where he has worked primarily on solid state devices and vacuum and gaseous electronics.

Dr. Webster is a member of Sigma Xi.

❖

V. M. Wolontis was born in Helsinki, Finland, in 1926. He received the M.A. degree in mathematics from the University of Helsinki in 1947 and the Ph.D. degree in mathematics from Harvard University in 1949. He was a teaching fellow at Harvard in 1948–49 and an assistant professor of mathematics at the University of Kansas from 1949 to 1953. In 1953 he joined the Mathematical Research Department of the Bell Telephone Laboratories.

V. M. Wolontis

Dr. Wolontis is a member of the American Mathematical Society and the Association for Computing Machinery.

❖

A. H. Zemanian (A'51) was born on April 16, 1925, in Bridgewater, Mass. He received the B.E.E. degree from the College of the City of New York in 1947 and the M.E.E. and Eng.Sc.D. degrees from New York University in 1949 and 1953, respectively. He taught in the electrical engineering department of C.C.N.Y. for the academic year 1947–48, and then joined the Maintenance Company, New York, N. Y. Since 1952 he has been teaching electrical engineering at New York University, where he is an assistant professor.

A. H. Zemanian

Dr. Zemanian is a member of the National Society of Professional Engineers, the American Institute of Electrical Engineers, Tau Beta Pi, Eta Kappa Nu, and Sigma Phi Omega.

1955 IRE National Convention Program

Waldorf-Astoria Hotel, Kingsbridge Armory, March 21–24, New York, N.Y.

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Registration

Members and visitors may register at either the Waldorf-Astoria Hotel (49th Street and Lexington Avenue) or Kingsbridge Armory (Kingsbridge Road and Jerome Avenue, Bronx) during the following hours:

Waldorf-Astoria Hotel

Monday	9 A.M.–5:30 P.M.
Tuesday	9 A.M.–8:00 P.M.
Wednesday	9 A.M.–6:00 P.M.
Thursday	9 A.M.–2:30 P.M.

Kingsbridge Armory

Monday	10 A.M.–10 P.M.
Tuesday	10 A.M.–10 P.M.
Wednesday	10 A.M.– 5 P.M.
Thursday	10 A.M.–10 P.M.

Technical Sessions

Over 250 technical papers will be presented in 55 sessions. A schedule of sessions and locations is given on page 348. Complete program of papers, authors, and 100-word summaries appears on pages 349–377.

Exhibits

The Radio Engineering Show, featuring 704 exhibits, will be held at the Kingsbridge Armory and at the Kingsbridge Palace, located one and a half blocks south of the Armory on Jerome Ave. Exhibit hours are from 10 A.M. to 10 P.M. every day but Wednesday, when they close at 5 P.M.

Transportation

Free bus service will be provided between the Waldorf-Astoria Hotel and Kingsbridge Armory. In addition, both locations are convenient to the Jerome Avenue line of the Lexington Avenue IRT subway.

Annual Meeting

Of particular interest to members will be the Annual Meeting of the IRE at the Waldorf, which opens the convention at 10:30 A.M. Monday, March 21. A. V. Loughren, Director of Research, Hazeltine Corporation, will be the principal speaker. The program will include reports from officers regarding IRE activities.

Convention Record

All available papers will be published in the 1955 IRE Convention Record. The Convention Record will be published in ten parts and will be available in June, 1955.

All persons who become paid members of a professional group by April 30, 1955 will automatically receive a free copy of that part of the Convention Record containing the papers sponsored by their professional group.

Convention Record parts may be ordered at the Convention Record Desk, Waldorf-Astoria, or from IRE headquarters.

Social Events

A Cocktail party will be held on Monday at the Waldorf-Astoria, the annual IRE Banquet will take place Wednesday evening at the same hotel, and a full program of women's activities is scheduled throughout the four days of the IRE Convention. For full details see News and Radio Notes section, page 379, of this issue.

1955 IRE CONVENTION PROGRAM

SCHEDULE OF TECHNICAL SESSIONS

	BELMONT-PLAZA	WALDORF-ASTORIA					KINGSBRIDGE ARMORY	
	Moderne Room	Starlight Roof	Astor Gallery	Jade Room	Sert Room	Grand Ballroom	Marconi Hall	Faraday Hall
Monday March 21 2:30 P.M.–5:00 P.M.	*Session 1* INSTRUMENTATION I	*Session 2* ANTENNAS AND PROPAGATION I—Antennas	*Session 3* MOBILE COMMUNICATIONS	*Session 4* COMMUNICATIONS SYSTEMS	*Session 5* INDUSTRIAL ELECTRONICS	*Session 6* TELEMETRY AND REMOTE CONTROL I—*Symposium:* Some Problems Associated With Telemetering and Remote Control of a Space Station	*Session 7* CIRCUIT THEORY I—*Symposium:* Network Design	*Session 8* AUTOMATIC CONTROL I
Tuesday March 22 10:00 A.M.–12:30 P.M.	*Session 9* TELEMETRY AND REMOTE CONTROL II—Remote Control	*Session 10* ANTENNAS AND PROPAGATION II—Antennas	*Session 11* AERONAUTICAL AND NAVIGATIONAL ELECTRONICS I—Airborne Devices and Environment	*Session 12* BROADCAST TRANSMISSION SYSTEMS I—TV Broadcasting	*Session 13* AUDIO I—General	*Session 14* INFORMATION THEORY I (Ends at noon)	*Session 15* (Joint) INSTRUMENTATION AND TELEMETRY AND REMOTE CONTROL	*Session 16* ELECTRON DEVICES I—Tubes
Tuesday March 22 2:30 P.M.–5:00 P.M.	*Session 17* INSTRUMENTATION II	*Session 18* ENGINEERING MANAGEMENT I—*Panel Discussion:* Operations Research—A Tool of Engineering Management	*Session 19* AERONAUTICAL AND NAVIGATIONAL ELECTRONICS II—Radar and Aircraft Landing Aids	*Session 20* BROADCAST TRANSMISSION SYSTEMS II—Color Television	*Session 21* AUDIO II—*Symposium:* Music, High Fidelity, and the Listener		*Session 22* TELEMETRY AND REMOTE CONTROL III—Recent Telemetering Developments	*Session 23* ELECTRON DEVICES II—Microwave Tubes
Tuesday March 22 8:00 P.M.–10:30 P.M.		*Session 24* AUTOMATIC CONTROL II—Trends in Automatization of Procedures and Processes in Business and Industry					*Session 25* AUDIO III—*Seminar:* Magnetic Recording for the Engineer	
Wednesday March 23 10:00 A.M.–12:30 P.M.	*Session 26* ULTRASONICS I	*Session 27* ELECTRONIC COMPUTERS I	*Session 28* MICROWAVE THEORY AND TECHNIQUES I—Microwave Components	*Session 29* PRODUCTION TECHNIQUES—Electronic Equipment Assembly Methods	*Session 30* (Joint) INSTRUMENTATION AND NUCLEAR SCIENCE	*Session 31* (Joint) SYMPOSIUM ON SPURIOUS RADIATION (Ends at noon)	*Session 32* CIRCUIT THEORY II—General Theory	*Session 33* ANTENNAS AND PROPAGATION III—*Panel Discussion:* Extended Range VHF and UHF Propagation
Wednesday March 23 2:30 P.M.–5:00 P.M.	*Session 34* ULTRASONICS II	*Session 35* ELECTRONIC COMPUTERS II—*Symposium:* The Design of Machines to Simulate the Behavior of the Human Brain	*Session 36* MICROWAVE THEORY AND TECHNIQUES II—Microwave Techniques	*Session 37* QUALITY CONTROL AND RELIABILITY STUDIES OF ELECTRONIC TUBES AND SYSTEMS	*Session 38* BROADCAST AND TV RECEIVERS		*Session 39* CIRCUIT THEORY III—Filters and Lines	*Session 40* ANTENNAS AND PROPAGATION IV—Propagation
Thursday March 24 10:00 A.M.–12:30 P.M.	*Session 41* MEDICAL ELECTRONICS I—Panel Discussion	*Session 42* INFORMATION THEORY II	*Session 43* ELECTRON DEVICES III—Cathode-Ray Type Tubes	*Session 44* COMPONENT PARTS I—Electro-Magnetic Devices	*Session 45* NUCLEAR SCIENCE I	*Session 46* ENGINEERING MANAGEMENT II—General (Ends at noon)	*Session 47* MICROWAVE THEORY AND TECHNIQUES III—Ferrites	*Session 48* ELECTRONIC COMPUTERS III
Thursday March 24 2:30 P.M.–5:00 P.M.	*Session 49* MEDICAL ELECTRONICS II—General	*Session 50* ENGINEERING MANAGEMENT III—*Symposium:* Management Selection as Viewed by Psychologists and Engineering Executives	*Session 51* ELECTRON DEVICES IV—Transistors	*Session 52* COMPONENT PARTS II—General	*Session 53* INFORMATION THEORY III		*Session 54* (Joint) MICROWAVE COMMUNICATIONS AND SYSTEMS	*Session 55* AERONAUTICAL AND NAVIGATIONAL ELECTRONICS III—Navigation

SUMMARIES OF TECHNICAL PAPERS

SESSION I

MON. 2:30–5:00 P.M.

BELMONT-PLAZA MODERNE ROOM

Instrumentation I

Chairman: P. S. Christaldi, *Allen B. Dumont Labs., Inc., Clifton, N. J.*

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1.1 "Direct-Reading Instrument for the Measurement of RMS Pulse Jitter," by *Jesse J. Taub and Charles I. Smith, Material Laboratory, New York Naval Shipyard, Brooklyn 1, N. Y.*

An instrument for the measurement of rms time jitter in the leading edges and trailing edges of video pulses is described. In conjunction with a detector, this device will also measure jitter in rf pulse trains. The instrument operates on the principle that a portion of the electrical energy of the jitter may be isolated from the energy of the pulse train by means of a low-pass filter. It can be shown mathematically that the portion of the jitter energy thus isolated can be converted to a voltage which is directly proportional to the rms time jitter in microseconds. This equipment is easily calibrated and is direct-reading. The jitter may have sinusoidal, random, or other waveforms.

1.2 "An Automatic Sonic Spectrum Analyzer and Curve Tracer," by *Edward F. Feldman, Panoramic Radio Products, Inc., Mt. Vernon, N. Y.*

The instrument described is an all-electronic panoramic spectrum analyzer which provides a means for rapid sonic Fourier analysis. A wide-band 40–20,000 cps logarithmic and three variable center-frequency linear-sweep modes are selectable by means of a switch. Both linear and two decade logarithmic vertical deflection amplitude axes are provided. The instrument's residual distortion is at least -60 db.

The crux of the log sweep design lies in the relationship between df/dt, the rate of scan, and the maximum achievable adjacent frequency component resolution. An expression is derived which relates the IF bandwidth to the instantaneous df/dt. By means of a dual swept crystal filter, the bandwidth is varied in synchronism with the log frequency scan to maintain optimum resolution at all frequencies.

A simple adjunct equipment converts the analyzer into an audio sweep generator and frequency response curve tracer which has unique properties. "Stopbands" more than 60 db down are measured accurately because of the synchronous detection action.

1.3 "A Simplified Method for the Measurement of Highly Linear Sawtooth Waveforms," by *Sherwood King, University of Tennessee, Knoxville, Tenn.*

A resistance-capacitance differentiating circuit is described which develops a constant amplitude signal during the useful portion of the sawtooth, with the departure from linearity appearing as deviations from the constant level. Following removal of the constant level, the deviations are amplified and displayed on a standard cathode-ray oscilloscope. In this way the effect of altering various circuit parameters in the sweep generator is immediately observed.

The equation necessary to determine the error arising from the use of an imperfect differentiator is derived, and an example of the successful use of this technique is given as well as the conditions necessary for applying this method to sawtooth waveforms of other amplitudes and periodicity.

1.4 "The Diagraph—A Direct-Reading Instrument for Graphic Presentation of Complex Impedances and Admittances," by *Richard C. Hess, Federal Telephone & Radio Co., Clifton, N. J.*

The necessity for measuring complex impedances at rf rapidly and accurately is becoming an increasingly important problem. An instrument is described which measures complex impedances and presents the results graphically in the most useful form (i.e., either on a Smith chart, Carter chart, or a Polar chart in terms of a complex reflection coefficient.) Use is made of directional couplers to measure the amplitude of the reflection factor directly and to project its magnitude on the chart by means of a light-spot galvanometer. The phase angle is determined semi-automatically and is indicated by a rotation of the chart. In addition to measuring impedances, the unusual flexibility of the device allows it to measure attenuation and phase shift in a four-terminal network and also the phase difference between two signals. This paper describes work done by the firm of Rohde and Schwarz, Germany.

1.5 "Measurement of Parameters that Determine Front Edge Response of Pulse Transformers," by *Isidore Bady, Signal Corps Engineering Labs., Ft. Monmouth, N. J.*

The equivalent circuit that is commonly used in the analysis of the front edge response of pulse transformers consists of the leakage inductance and distributed capacitance of the pulse transformer. Several methods that have been used to measure these parameters are analyzed and sources of errors discussed. A novel method for the measurement of the parameters is described. Results obtained with this method agree very well with theoretically calculated values. A large amount of test data on many different type transformers is presented to support the theoretical analysis.

SESSION II

MON. 2:30–5:00 P.M.

WALDORF-ASTORIA STARLIGHT ROOF

Antennas and Propagation I Antennas

Chairman: John V. N. Granger, *Stanford Research Institute, Stanford, Calif.*

2.1 "Efficiency of Surface Wave Excitation," by *Alan F. Kay, Technical Research Group, New York, N. Y., and Francis J. Zucker, Air Force Cambridge Research Center, Cambridge, Mass.*

The excitation efficiency of surface waves is the ratio between the power in an open-wave-guide mode and the total power available at the feed. It should be maximized if the open wave-guide is used as a transmission line; and it should be a predictable quantity if the guide is used as an antenna whose pattern is to be modified in some specified manner by the feed radiation.

A method based on Fourier transforms is used to calculate the excitation of dipoles, slots, and extended sources of arbitrary orientation. Some of the predictions have been verified experimentally. The question of optimum efficiency obtainable for diverse surface wave applications is also discussed.

2.2 "Serrated Waveguide: Theory and Experiment," by *R. S. Elliott and K. C. Kelly, Hughes Research and Development Lab., Culver City, Calif.*

An experimental study of linear arrays of closely spaced nonresonant transverse slots in X-band waveguide is described and analytical expressions for the complex propagation constant in the guide proper are given. These closely slotted (or serrated) waveguides have the advantage of broadband operation and may be used as flush-mounted antennas in many applications. Normally the beam maximum occurs at a tilt angle from endfire equal to $\cos^{-1} c/v_g$. The guide velocity, v_g, may be altered by the use of a ridge in the guide, or a dielectric, or by varying the thickness of the slotted wall. The last mentioned method has the ability to speed or slow the wave in the guide depending on the wall thickness and the slot length. Aperture illumination can be controlled by varying slot length and spacing. Characteristics obtained and applications attempted are given in some detail, and the wide range of possible beam shapes is discussed.

2.3 "Properties of a Radiating Discontinuity on a Corrugated Surface Transmission Line," by *M. J. Ehrlich and I. K. Williams, Microwave Radiation Co., Inc., Venice, Calif.*

Although much work has been done, both theoretically and experimentally, on surface wave transmission lines and surface wave radiators, no data have been available regarding the radiation properties and equivalent network representation of radiating discontinuity on the surface wave transmission line.

An experiment was designed to obtain the free-space radiation patterns and equivalent network representation of a discontinuity on a corrugated surface-wave transmission line as a function of the parameters of the discontinuity and line.

Data obtained were then used to formulate an empirical theory which furnishes the four-terminal network representation and free-space radiation characteristics.

These data and the empirical theory allow the design of shaped beam antennas, physically realized by surface-wave transmission lines utilizing corrugation discontinuities as the radiating elements, to be carried out by current linear array design theories.

2.4 "Symmetrical Microwave Lenses," by *C. Goatley and C. F. Parker, Melpar, Inc., Alexandria, Va.*

Three new microwave lenses are described which have the common property that they focus energy arriving from two diametrically opposite directions in space. One of these lenses with constant index of refraction is related to the double convex lens of optics while the other two are formed of variable index of refraction media. The derivations of the equations for each lens are briefly discussed and methods of construction are indicated. Experimental models of each lens are described together with results of radiation pattern measurements.

2.5 "Lens and Feed System for Volumetric Scanning GCA Antenna," by *George D. M. Peeler and William F. Gabriel, Naval Research Lab., Washington, D. C.*

An *X*-Band, bifocal, constrained, metal plate lens, 10×15 feet, has been designed to produce a beam $\frac{1}{2}\times\frac{2}{3}$ degrees, and to permit scanning this beam over a volume $6\frac{2}{3}$ degrees in elevation by 20 degrees in azimuth (±6.7 beamwidths in elevation by ±13.3 beamwidths in azimuth). The correction points are at ±10 degrees azimuth, 0 degrees elevation. A plane focal surface is obtained by stepping the lens in approximately hyperbolic contours to correct phase errors on-axis while maintaining perfect focus at the correction points. Wavefront calculations show deviations less than one-tenth wavelength from a plane wave for all feed positions within the volume of interest.

The subject antenna was proposed by Naval Research Laboratory for obtaining a rapid, two-dimensional scan directly applicable to the GCA (Ground Control Approach) aircraft landing problem. It is to consist of a lens, a stacked organ pipe, and a ring scanner. An *X*-Band prototype sponsored by the Air Force, is now under development which will scan a sector of $6\frac{2}{3}$ degrees in elevation by 20 degrees in azimuth with four pencil beams of beamwidths $\frac{1}{2}$ degree in elevation by $\frac{2}{3}$ degree in azimuth. A "television" type scan action is to be employed in which there are 20 horizontal scan lines spaced $\frac{1}{3}$ degree apart and the frame repetition rate is 4 per second. To accommodate the four pencil beams, the $6\frac{2}{3}$ degrees elevation scan is to be divided into four sectors of $1\frac{2}{3}$ degrees, all four sectors to be scanned simultaneously. An experimental working model of the feed system (organ pipe plus ring scanner) has been built and tested, with satisfactory results obtained therefrom.

SESSION III

MON. 2:30–5:00 P.M.

WALDORF-ASTORIA ASTOR GALLERY

Mobile Communications

Chairman: DANIEL E. NOBLE, *Motorola, Inc., Chicago, Ill.*

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3.1 "An Experimental Mobile Dispatching System," by *R. W. Collins and V. A. Douglas, Bell Telephone Labs., New York, N. Y.*

This mobile telephone system provides dispatch service for a number of small vehicular groups and their respective dispatchers on one radio channel. Wire lines connect the dispatchers' standard manual telephone sets directly to common terminal and signaling equipment arranged to control the radio transmitter and receiver and to comply with FCC rules. The fleets use commercially available radio equipments modified to add suitable tone generating and receiving equipment. It is anticipated that the system will be used for large numbers of brief calls and will increase considerably the amount of dispatch type traffic that can be handled over telephone company mobile telephone facilities.

3.2 "450 MC Mobile Equipment Employing Direct Frequency Modulation," by *W. Ornstein, Canadian Marconi Co., Montreal, P.Q., Canada.*

A 450 mc mobile radiotelephone having several unique features is described. The transmitter oscillator stage has an output frequency of 150 mc and employs a 50 mc overtone crystal. Direct frequency modulation of the crystal is accomplished using only passive elements and retaining a high order of frequency stability in the oscillator circuit.

The transmitter employs only four rf stages including the oscillator and has an output power level of 7.5 w.

The receiver is a double conversion superheterodyne employing a disc seal triode in a coaxial cavity as the input rf amplifier. Receiver noise figure is 10 db and sensitivity (on a 12 db signal to noise basis) is 0.5 microvolts.

3.3 "Design Problems of VHF Repeater Stations," by *J. R. Neubauer, Radio Corp. of America, Camden, N. J.*

This is a discussion of the physical and electrical design considerations which determine the functional effectiveness and operating reliability of stations employing simultaneous operation of transmitter and receiver as a control or reradiating system. Consideration will be given to requirements for station location, primary and emergency power control and regulation, antennae supports and radio equipment control and supervision. In addition, the effects of simultaneous operation on the system performance will be evaluated. Methods of overcoming the effects of desensitization, intermodulation, spurious emission and response, distortion and noise are discussed. The importance of adequate study of propagation and frequency allocations are emphasized. Slides will be used to illustrate pertinent circuit considerations and interesting applications.

3.4 "Evaluation of Sideband Noise and Modulation Splatter of VHF Transmitters," by *W. L. Firestone, Motorola, Inc., Chicago, Ill.*

The ever-increasing demand for additional spectra in the vhf and uhf bands emphasizes the problems of transmitter sideband noise and modulation sideband splatter.

This paper is concerned with the determination and the reduction of the various factors which contribute to sideband splatter and noise. Splatter created as the byproduct of the modulation processes in typical AM, PM, and FM transmitters is analyzed, measured, and compared. Measurements of AM spectra show that this splatter exceeds the expectations based on simple theory and nonlinear theory is shown to predict the additional measured side bands. Noise sources in typical FM sets are determined and methods of reducing noise are discussed. The addition of a low-pass filter as the last element in the audio system is shown to be a must in both FM and AM if noise and sideband splatter are to be controlled.

3.5 "A Miniature Reflectometer for Portable and Mobile Transmitters," by *Edwin M. Stryker, Jr., Collins Radio Co., Cedar Rapids, Ia.*

It is desirable to incorporate into transmitters, especially those of the portable and mobile types, a reflectometer or standing-wave indicator to provide a means of determining transmitter output and proper termination and functioning of the antenna and its transmission line. The reflectometer has the advantages of small size and low cost as well as being useful over a wide frequency range. Employing printed circuitry and requiring no adjustment to permit directional coupling, this device lends itself well to mass production techniques. Principles applicable to this type of directional coupler are discussed and test data are given.

SESSION IV

MON. 2:30–5:00 P.M.

WALDORF-ASTORIA JADE ROOM

Communications Systems I

Chairman: F. M. RYAN, *American Telephone and Telegraph Co., New York, N. Y.*

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4.1 "A New Horizon in Communication Theory—The Polyphase Concept," by *Allan A. Kunze and John G. Schermerhorn, Rome Air Devel. Center, Griffiss Air Force Base, Rome, N. Y.*

Communication engineering has been treated, hitherto, generally from single-phase concepts only. For instance, the recognition of push-pull circuitry as a two-phase polyphase circuit has not been emphasized in communication theory. Since it is believed that the significance of the polyphase concept and its application is not known clearly, as yet, this paper attempts to fill such a need. It is demonstrated that the polyphase concept is of sufficient scope and power to penetrate into every domain constituting communication theory and provide new approaches to old problems—information theory, modulation, propagation, antennas, detection, tubes, circuitry and even acoustics. The Fortescue symmetrical component theory is borrowed from the power engineer and its extension is shown to be a powerful analytic tool in communications analysis.

Further outlined are the possibilities of new antennas, transformers, circuitry, instruments and generally new techniques for application in

communications resulting from preliminary considerations of polyphase applications. In particular, such applications as interference suppression, multiplex, high-power generation, and wide aperture D/F's are discussed.

4.2 "A Theorem Concerning Noise Figures," by *A. G. Bose and S. D. Pezaris, Massachusetts Institute of Technology, Cambridge, Mass.*

This theorem concerns the greatest lower bound of the noise figure of a system consisting of n identical tubes whose plate signals are directly added and whose inputs are obtained through a linear passive coupling network from a source having a finite impedance. It is proved that the optimum attainable noise figure of this system is equal to the optimum noise figure of a circuit consisting of one such tube and an appropriate coupling network. The theorem is extended to the lower bound of the noise figure of a similar circuit in which arbitrary phase shifts are introduced before the plate signals are added. This extension includes the use of distributed-line amplifiers.

4.3 "Automatic Operation of a High-Power Amplifier," by *V. R. DeLong, Collins Radio Co., Cedar Rapids, Iowa.*

Automatic operation of a high-powered radio frequency amplifier operating in the 1.5 to 30 mc frequency range will be described. Details of a unique switchless coarse-fine information system will be explained. Emphasis will also be placed upon the problems involved in automatically adjusting the tank circuit and load control of the amplifier when operating in single sideband service.

4.4 "The Use of Reflex Techniques in a VHF-UHF Communication System," by *Paul G. Wulfsberg, Collins Radio Co., Cedar Rapids, Iowa.*

The use of novel reflex techniques in the design of a new 1800 channel vhf-uhf communication set are described. Reduction of nearly 50 per cent in tube count and 25 per cent in tuned circuits is achieved in a transmitter-receiver unit employing direct frequency synthesizing techniques. A new transmitter circuit is described which eliminates the conventional sidestep oscillator and mixer. Methods permitting the transmit and receiver spurious responses to be suppressed nearly 80 db are outlined.

The use of the foregoing techniques permits design of a radio set of unusually small size and very high reliability achieved by a 40 per cent reduction in total electrical components.

4.5 "A New Teletypewriter Using the Integration Method of Detection," by *Henning F. Harmuth, Signal Corps Engineering Labs., Ft. Monmouth, N. J.*

In standard teletypewriters detection of the signals is performed by sampling pulses and pulse spaces. Instead of this, one may integrate in several ways over the signals and use the integrals obtained for the distinction of the different signals. This integration method is less susceptible to noise than the sampling method since the positive and negative noise amplitudes partly cancel each other. Some basic aspects of this integration method are discussed and its application to the construction of a teletype receiver is shown.

SESSION V

MON. 2:30–5:00 P.M.

WALDORF-ASTORIA SERT ROOM

Industrial Electronics

Chairmen: CONAN A. PRIEST, *Onondaga Pottery Co., Syracuse, N. Y., and* WILFRID L. ATWOOD, *Warren, Ohio*

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5.1 "An Instrument to Count and Size Particles Suspended in a Gas," by *Ernest S. Gordon, Illinois Institute of Technology, Chicago, Ill.*

An instrument embodying optical and electronic principles has been developed to count and size aerosol particles automatically. This device, known as an aerosoloscope, eliminates the tedious and time-consuming method employed by chemists in microscopic counting of particles collected on a slide. Among the applications of the aerosoloscope are the study of air pollution problems, milling operations, and precipitation monitoring at weather stations.

Particles from 1 to 64 microns in diameter and counting rates up to 150 particles per second are handled by the instrument described. The particles are diluted and drawn across a high intensity beam of light; a multiplier phototube receives the light scattered by each particle at 90 degrees. The resulting impulses are amplified and separated into 12 size groups by a unique pulse height discriminator. Glow transfer decade counters are used to count the pulses in the various size groups.

5.2 "Design Considerations of Microwave Ovens," by *Robert A. Rapuano and Robert V. Smith, Raytheon Manufacturing Co., Newton, Mass.*

Practical electronic ovens offering the desirable characteristics of food cooked with microwaves have been developed using a reliable high power cw magnetron at 2,450 mc.

Food has the properties of a lossy dielectric, and therefore can be cooked by the heat resulting from the absorption of microwaves in a multi-resonant cavity with a microwave efficiency greater than 90 per cent.

Practical designs require that the microwave energy distribution be uniform and that the resonant cavity be the proper rf load for the magnetron over a wide range of food loading.

Methods are developed for determination of proper magnetron operation into various cavity designs. Criteria for an isotropic pattern oven are given.

5.3 "High-Speed Electronic Fault Protection for Power Tubes and Their Circuitry," by *W. N. Parker and M. V. Hoover, Radio Corp. of America, Lancaster, Pa.*

High-speed electronic circuits have been developed to minimize the possibility of damage to power tubes as a consequence of the "flash-arc" or "Rocky Point Effect." The circuit detects the development of fault conditions in a power tube (and/or its circuitry) and triggers a gaseous conduction device connected in shunt with the dc power supply. Fault currents are rapidly bypassed from the faulting power tube by means of the gaseous conduction device; consequently, the "flash-arc" in the protected power tube is extinguished before serious damage results. The gaseous conduction device bypasses the rectifier-output and filter-circuit energy until such time as the rectifier can be de-energized.

5.4 "A Magnetic Thyratron Grid Control Circuit," by *James H. Burnett, Electrons, Inc., Newark, N. J.*

A new compact magnetic method for driving thyratron grids has recently simplified high-power fast response servos and regulated power supplies.

Since large power gain is available in the thyratron, power amplification can be sacrificed in the magnetic circuit to gain speed of response comparable to the thyratron inverse cycle. The new grid circuit provides a rapid rate of rise of thyratron firing voltage, wide-control range and high-voltage amplification. Dc or ac input signals to the magnetic circuit work equally well, and the signal voltage source can be isolated from the power circuit.

5.5 "A Static Frequency Detector, Magnetic Type," by *Henry W. Patton, Collins Radio Co., Cedar Rapids, Iowa.*

A static frequency detector operable at audio frequencies will be described. This detector circuit delivers a dc output signal accurately and linearly proportional to input frequency. Large variations in input signal amplitude, signal distortion, and environmental conditions do not change the output signal amplitude. It is easy to construct, inexpensive, and when once calibrated has an almost indefinite life. The circuit is simple, compact, and does not require a power supply. Input signal energy requirements are less than a watt. The circuit has application to audio frequency measurements, quartz crystal manufacture, FM telemetering, and tachometers.

5.6 "A New Machine for Automatic Production of Electronic Assemblies," by *Cledo Brunetti, T. R. James, C. H. Bergsland, and D. F. Melton, General Mills, Inc., Minneapolis, Minn.*

SESSION VI

MON. 2:30–5:00 P.M.

WALDORF-ASTORIA GRAND BALLROOM

Telemetry and Remote Control I Symposium: Some Problems Associated with Telemetering and Remote Control of a Space Station

Chairman: J. GORDON VAETH, *Office of Naval Research, Port Washington, N. Y.*

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This is a symposium which will discuss some of the problems associated with telemetering and remote control of a space station. *Problems*

to be discussed include: *problems* associated with acquisition of data on the nature of space; *problems* associated with an ionic rocket engine; *problems* associated with construction of a space platform; *problems* associated with the tracking of a space station; *problems* associated with training of men to operate in space; *problems* associated with telemetering and communication.

It is the purpose of this symposium to acquaint the audience of Radio Engineers with some of the basic problems with which he may some day be faced.

6.1 "The Use of Piloted Balloons as Space Laboratories," by *M. D. Ross, Office of Naval Research, Washington, D. C.*

6.2 "Ionic and Nuclear Problems of Rocket Propulsion," by *F. J. Murray, Columbia University. New York, N. Y.*

6.3 "Instrumentation of a Minimum Satellite for Astro-Physical Research," by *S. F. Singer, University of Maryland, Baltimore, Md.*

6.4 "Telemetering and Control of a Space Station," by *Wernher Von Braun, Redstone Arsenal, Huntsville, Ala.*

6.5 "Telemetering in the Development of Space Flight," by *C. B. Ruckstuhl, Jr., Bendix Aviation Corp., N. Hollywood, Calif.*

6.6 "Synthetic Training for Space Flight," by *G. V. Amico, Office of Naval Research, Port Washington, N. Y.*

SESSION VII

MON. 2:30–5:00 P.M.

KINGSBRIDGE ARMORY
MARCONI HALL

Circuit Theory I
Symposium: Network Design

Chairman: CHESTER H. PAGE, *National Bureau of Standards, Washington, D. C.*

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7.1 "Influence of Computing Machines on Network Design Methods," by *John T. Bangert, Bell Telephone Labs., Murray Hill, N. J.*

For years network design has relied on a combination of theory and experience, supplemented by calculation. With this procedure calculation has played an important though subordinate part. The spectacular development of electronic computers has provided cogent reasons for revising certain design techniques to give computation a major role. This has stimulated the evolution of a new design philosophy which will be illustrated by a group of selected examples. These involve such topics as iteration, conformal mapping, and the potential analog.

In this new philosophy the computer serves two major functions. The first is to uncover any promising solutions by rapid exploration of many areas. The second is to perform extensive mathematical operations on large quantities of numerical data with great speed and with complete freedom from error. The use of modern computers has made the design of networks of extreme precision and complexity economically possible. In addition certain intractable problems formerly treated empirically are gradually being reduced analytically.

7.2 "The Use of Potential Analogs in Network Synthesis," by *Ronald E. Scott, Massachusetts Institute of Technology, Cambridge, Mass.*

A summary is given of the use of potential analogs for synthesis in the time domain, the frequency domain, and the real-part domain. Most of the analogs discussed are based on current flow in a resistive sheet. In addition a new device is described which combines the three types and yields essentially instantaneous curves of the gain function, the phase function, the real part, the imaginary part, and the transient response. This device can be used for network synthesis by any of the three methods and also for checking the relations between them. Throughout the discussion particular attention is devoted to the use of symmetry and conformal mapping to reduce the size of physical devices and decrease the inherent errors.

7.3 "Iterative Network Synthesis," by *Howard B. Demuth, Stanford Research Institute, Stanford, Calif., and George A. Caryotakis, A. Donald Moore, and David F. Tuttle, Stanford University, Stanford, Calif.*

An iterative method of network synthesis, believed to be new, was developed at Stanford by A. D. Moore in 1952. The method was originally applied to distributed amplifier design, but recently it has been used to synthesize low-pass stagger-tuned amplifiers having special amplitude or phase characteristics.

The iterative method consists of alternately satisfying two relationships between the poles and zeros of the system function. The first relationship R_1 is obtained by specifying the network configuration. Given the network configuration, a set of zeros may be obtained from a set of poles. A realizable amplitude or phase restriction on the system function yields a second relationship R_2, which may be used to determine a set of poles from the zeros. Maximally-flat or equal-ripple amplitude restrictions, or a linear phase restriction, have been used to produce R_2. In each step of the iteration process one obtains a set of zeros from a set of poles using R_1, and then one obtains a new set of poles from these zeros using R_2. The object is to find poles and zeros which satisfy both R_1 and R_2, and thereby determine a network with the required configuration which does have the desired phase or amplitude characteristic. This iterative process has been found to converge rapidly enough to be useful. As factorization of high-order polynomials, or evaluation of large determinants, is required in each cycle of iteration, use of a digital computer is necessary.

The designs produced by this method make best use of the available degrees of freedom. An optimum amplitude or phase characteristic can be attained without resorting to cancellation of the system function zeros by use of extraneous poles. Pole-zero cancellation has been used in the past to produce an all-pole function. In the absence of zeros the location of the poles to produce a certain characteristic is generally known. However, pole-zero cancellation wastes degrees of freedom and one fails to attain optimum gain-bandwidth or phase linearity, as the case may be.

An example of iterative amplifier synthesis will be described and other designs will be presented and compared with conventional amplifiers.

7.4 "The Use of Least Squares in Network Design," by *M. R. Aaron, Bell Telephone Labs., Murray Hill, N. J.*

Least squares is an effective tool in performing the various steps in network design from approximation to network alignment. The common denominator in each of these procedures is the approximation of a desired network performance or change in performance by a linear sum of functions. After this preconditioning, the mathematical steps that remain involve matrix multiplication and the solution of a set of linear simultaneous equations (matrix inversion). Both of these processes have been programmed for solution by a digital computer. The marriage of digital computers to this means of network design has been highly compatible. Several "offspring" are displayed to attest to this high degree of marital bliss. These examples include: (1) Alignment of a one-port to match a prescribed magnitude characteristic; (2) Solution of the approximation problem in the design of a two-port; (3) Synthesis of all-pass networks into a composite delay equalizer.

7.5 "Summation and Outlook," by *Ernest A. Guillemin, Massachusetts Institute of Technology, Cambridge, Mass.*

SESSION VIII

MON. 2:30–5:00 P.M.

KINGSBRIDGE ARMORY
FARADAY HALL

Automatic Control I

Chairman: JOHN E. WARD, *Massachusetts Institute of Technology, Cambridge, Mass.*

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8.1 "Analysis of Combined Sampled- and Continuous-Data Systems on an Electronic Analog Computer," by *Louis B. Wadel, Chance Vought Aircraft, Inc., Dallas, Texas.*

Combination sampled- and continuous-data dynamic systems can be studied on a conventional electronic analog computer. The only auxiliary equipment needed is a timing device

to control relays already contained in the computer. Each combination of sampling-switch and zero-order hold device is simulated by one integrator connected as a store and controlled in state separately from the integrators in the continuous-data portion of the circuit. Simulation of higher-order hold devices require additional storage integrators. Real time solution is not obtained because computer time halts during sampling.

8.2 "An Adaptive Servo System," by *A. H. Benner and R. F. Drenick, Radio Corp. of America, Camden, N. J.*

An adaptive servo system is here defined as a servo which can adjust its parameters in accordance with the character of its input. It is intended to follow chiefly inputs which vary linearly with time. It is therefore designed for this purpose as a linear device. However, on occasion the signal is expected to vary quadratically with time. This suggests that an adaptive feature of the servo be added which allows it to change its parameters for tight and fast follow-up. A linear system is first described and an equation for signal prediction is derived. The limits of stability of this system are shown graphically and analyzed. It is then shown that the conditions for optimum performance for following a quadratic signal are strictly incompatible with those for a linear one.

The nature of the adaptive system is discussed and the necessary limiting conditions for its use are stated. It is shown that the decision as to when to shift to the tight mode of operation and back, can be based on the instantaneous follow-up error. Optimum decision rules are drawn up and discussed. Finally the performance of an adaptive servo is compared to a linear servo for typical statistical and nonstatistical inputs. Considerable improvement in performance is indicated.

8.3 "Application of a Magnetic Amplifier to a High-Performance Instrument Servo," by *Paul R. Johannessen, Massachusetts Institute of Technology, Cambridge, Mass.*

The design of a high-performance two-phase motor servomechanism utilizing a magnetic amplifier as the power stage will be described. For compensation, the servomechanism employs tachometer feedback, passed through a high-pass filter to obtain a high velocity constant. The magnetic amplifier not only provides a convenient, compact power output stage, but also has the advantages of being insensitive to quadrature and of being able to accept as its input the sum of an ac error signal and a dc tachometer signal. The high-pass tachometer filter blocks the transmission of dc bias voltages to the magnetic amplifier, so that the servomechanism has extremely low drift.

8.4 "A Nonlinear Compensating Configuration for Saturating Servomechanisms," by *W. H. Surber, Jr., Princeton University, Princeton, N.J.*

A nonlinear compensating circuit configuration for a feedback control system which can appreciably improve its performance relative to that obtainable using conventional types of control networks is analyzed and the system operating characteristics studied. Experimental results are given for an actual positional servomechanism and for several somewhat idealized systems simulated with an analog computer.

Two general classes of nonlinearities are important in the operation of high performance servomechanisms. The ultimate obtainable dynamic performance is limited by saturation in the power amplifying elements and in the output energy conversion device. For optimum utilization of equipment, the output device should be operated at its maximum or saturation level for relatively small errors or error derivatives. In the neighborhood of the error null, however, a group of "small signal" nonlinearities such as potentiometer resolution, backlash, coulomb friction and others impose limitations on the system performance and affect the control network design in order to prevent small amplitude hunting.

A nonlinear auxiliary feedback loop consisting of a regenerative connection of the output rate signal through a nonlinear network and a stabilizing high-pass filter is shown to provide a close approach to the optimum recovery of the system from saturation without affecting the error null stability as is the case for certain other types of nonlinear control configurations for saturating servomechanisms. This can be accomplished with relatively simple and reliable networks.

8.5 "Delay-Line Methods for Compensating Closed-Loop Systems in the Time Domain," by *R. E. Scott and Y. C. Ho. Massachusetts Institute of Technology, Cambridge, Mass.*

Theoretical, analog, and experimental results are given for a new method of compensating feedback networks in the time domain. The method is based upon the open-loop impulse response of the system, and it yields the desired closed-loop transient response and the required compensation network. The compensation network is obtained in the form of a distributed element transfer function which can be realized approximately by lumped parameter elements. If the design criteria are given in the time domain, a simple iterative procedure will lead to the desired result. If the design criteria are given in the frequency domain, the required network is obtained by the inversion of a specified matrix.

SESSION IX

TUES. 10:00 A.M.–12:30 P.M.

BELMONT-PLAZA MODERNE ROOM

Telemetry and Remote Control II Remote Control

Chairman: CHARLES H. DOERSAM, JR., *Sperry Gyroscope Co., Great Neck, N. Y.*

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9.1 "New Apparatus and Techniques of Air Traffic Control Data Handling and Display," by *David J. Anthony, Convair, San Diego, Calif.*

General limitations of conventional radar identification techniques are defined insofar as they place limits upon the information capacity of compatible codes. A complete system is described wherein a new type of display tube is utilized. This tube produces joint displays of radar data and identification symbols wherein symbol displays are directly formed without the necessity for area scanning. An associated computer and control system is described, which serves to convert time sequential codes into analog data for control of the tube display. Details of the display tube, the manner in which registration difficulties are minimized, code- and timing-sequences are discussed.

9.2 "The Role of the Digital Computer Processing in Guided Missile Data," by *H. N. Morris, RCA Missile Test Proj., Patrick AF Base, Fla.*

The high-speed automatic digital computer has become an indispensable tool in the processing of guided missile data. Major utilization has been in the field of optical position and attitude data reduction. One example is the computing of position of a vehicle in space, given azimuth, elevation, and timing information from two or more instruments, along with accurate position location of the data-gathering cameras. Electronic position and trajectory information provide an ever-increasing workload for the computer as new type radar and other electronic measuring devices are placed into operation. Last, but certainly not destined to be least, is the field of telemetering. The highly versatile digital computer offers tremendous possibilities in the processing of this "internal" data so as to increase by large factors both the speed and accuracy of the reduction.

9.3 "A New Method for Designing the Compensation of Feedback Control Systems," by *Gilbert S. Stubbs, The Franklin Institute, Philadelphia, Pa.*

It is often convenient to separate the synthesis of the open-loop transfer function of a feedback control system from the details of compensation circuitry. A method for accomplishing this end is outlined in this paper. A significant feature of the method is the representation of the control compensation by a single-series transfer function equal to the system open-loop function, divided by the product of the transfer functions of the fixed components in the open loop. A few general restrictions on this series-equivalent compensation function usually suffice to insure realizable compensation circuits. Graphical and analytical methods for synthesizing the compensation circuit functions to obtain the series-equivalent function are described.

9.4 "Analysis of Sampled Data Systems and Digital Computers in the Frequency Domain," by *Rubin Boxer, Rome Air Devel. Center, Griffiss AF Base, Rome, N. Y.*

Functions defined at discrete intervals, i.e., sampling or impulse functions are analyzed. These are shown to be Laplace transformable utilizing either jump or impulse functions. The relation between jump function transforms and z-transforms is derived.

Methods for solving linear difference equations utilizing z-transforms are introduced. The particular difference equations utilized as examples are derived from digital positional and velocity-aided track with scan-feedback loops.

The concept of transfer function for sampled data systems is utilized. Equivalent Nyquist diagrams are derived for a frequency analysis of the systems previously solved by the difference equation method.

The utility of the techniques for analyzing digital computer programs is illustrated by a frequency analysis of numerical analysis methods of differentiation and integration.

Limitations of the methods are included.

SESSION X

TUES. 10:00 A.M.–12:30 P.M.

WALDORF-ASTORIA STARLIGHT ROOF

Antennas and Propagation II Antennas

Chairman: GEORGE SINCLAIR, *Sinclair Radio Labs., Ltd., Ontario, Can.*

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10.1 "Omnidirectional Circularly Polarized Antennas," by *K. S. Kelleher and C. W. Morrow, Melpar, Inc., Alexandria, Va.*

This paper considers several circularly polarized antennas with radiation patterns which are omnidirectional in azimuth and which, in the vertical plane have, respectively, a broad beam, a narrow beam and a shaped beam. The broad-beam pattern is obtained from a biconical horn structure in which circular polarization has been introduced by means of depolarizing elements. The other two antennas consist of toroidal lenses fed by the biconical horn radiator. Because of the symmetry of these lenses, their design can be obtained from a two-dimensional analysis. For the narrow beam lens, the analysis is straightforward. The shaped-beam system required a novel design.

10.2 "The N.R.L. Precision 'Big Dish' Antenna," by *D. L. Holzschuh, Collins Radio Co., Cedar Rapids, Iowa.*

The highest gain antenna in the world today is the 50-foot parabolic reflector built for the Naval Research Laboratory for operation at wave-lengths as short as 8 mm. A description of this precision antenna, which is used for research in radio astronomy, is presented. The mechanical aspects of the dish are given primary consideration. Fabrication procedures, design considerations, and a description of the unique surface machining apparatus are treated—these techniques yielding a reflector with an over-all surface accuracy of ± 0.020 inches. A mention is made of a few of the recent advances in radio astronomy made at N.R.L. through the use of the Big Dish—testifying to the successful completion of the project.

10.3 "The Omniguide Antenna; an Omnidirectional Waveguide Array for UHF-TV Broadcasting," by *O. M. Woodward, Jr., RCA Laboratories, Princeton, N. J., and James Gibson, Vattersnas, Sweden.*

The Omniguide antenna is a new type of high-gain antenna developed for UHF-TV broadcasting. Waveguide components are employed in the design instead of coaxial line elements in order to increase the power-handling capacity. Separate picture and sound inputs are provided which are decoupled independent of frequency, thus eliminating the need for a frequency-selective combining filter. A reflection-absorbing circuit increases the picture input bandwidth with very small sacrifice of power.

Experimental measurements taken on an antenna of this type constructed for UHF Channel No. 72 are described. Considerations are given to several alternative variations of the system to cover the entire UHF range.

10.4 "The Circular Traveling-Wave Antenna," by *W. J. Bergman and F. V. Schultz, The University of Tennessee, Knoxville, Tenn.*

This antenna consists of a wire formed into a circle, fed at one point and so terminated at the diametrically opposite point that only a traveling wave exists on the antenna. The azimuthal radiation patterns have been calculated over a frequency range of three-to-one, and measured over a frequency range of eight-to-one. The patterns have a single main lobe with rather high-level sidelobes. An initial investigation has been made of the possibility of reducing the sidelobe level by using two concentric circular antennas having different radii.

10.5 "Stripline Radiators," *James A. McDonough, Airborne Instruments Lab., Inc., Mineola, N. Y.*

The application of etching and printing techniques to electronic circuits makes feasible the construction of microwave strip antenna arrays both from practical and economical viewpoints. Many conventional broadside curtains used at lower frequencies were impractical to fabricate in the region above 1,000 mc. Progress in the design of strip conductors as microwave antennas indicates that many of these arrays can now be adapted for operation at microwaves and fabricated with sufficient accuracy to be truly useful. Some of the components involved in the design of strip antennas, such as balanced lines, twists, baluns, and power dividers have been investigated. Complete strip arrays consisting of up to 80 elements have been tested. The gain, bandwidth, and radiation patterns appear comparable to the results obtained from the corresponding arrays used at the lower frequencies for which the same theory of operation applies.

SESSION XI

TUES. 10:00 A.M.–12:30 P.M.

WALDORF-ASTORIA ASTOR GALLERY

Aeronautical and Navigational Electronics I Airborne Devices and Environment

Chairman: G. L. HALLER, *General Electric Co., Syracuse, N. Y.*

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11.1 "Aircraft Electronics—Environment, Specifications and Survival," by *F. Mintz and M. B. Levine, Armour Research Foundation, Chicago, Ill.*

The widespread use of electronics in military aircraft requires constant evaluation of the design criteria as related to reliability. Of considerable importance are the environmental factors: mechanical shock, vibration, and noise. This paper examines the existing environments as measured aboard both reciprocating and jet engine aircraft, and draws a comparison with current military specifications governing shock and vibration testing of electronic components. In addition, the results of shock and vibration testing on representative airborne electronic components are presented. Also included are the effects of high-intensity sound on certain components. Some recent work shows that high-intensity acoustic excitation can produce malfunctions of electronic components, although no government specifications cover acoustic effects. Much of the work discussed was sponsored by the Wright Air Development Center.

11.2 "Dynamic Environmental Testing of Airborne Electronic Components," by *R. H. Jacobson and M. B. Levine, Armour Research Foundation, Chicago, Ill.*

The effects of vibration, shock, and high intensity acoustic excitation on airborne electronic components and equipments were investigated. The investigation was conducted under Air Force sponsorship with the ultimate objective of enhancing the reliability of electronic equipment utilized in conventional aircraft as well as missiles.

Tubes, relays, potentiometers, transformers, fixed capacitors, and pressure switches were included among the components which were studied. The paper describes the techniques and procedures followed and summarizes the test results obtained.

11.3 "A Communication Theory Approach toward the Design of Aircraft Instrument Displays," by *Lawrence J. Fogel, Stavid Engineering, Inc., Plainfield, N. J.*

The effectiveness of a military aircraft depends upon the transmission of information from the instrument display to the pilot. In the past, this communication link has received much empirical study. Now, however, an analytical approach is possible which will not only indicate which display design is better, but will also provide criteria to insure future displays that *will* meet the specified requirements of the aircraft system wherein they must operate. An approach-strategy for such a research program is presented in this paper, together with its limitations and an indication of those aspects which must yet be studied before the final solutions will be achieved. Although much remains to be done, sufficient data are currently obtainable to allow displays to be designed with the knowledge that they will fulfill their mission.

11.4 "The C19K Charactron," by *J. T. McNaney, Convair, San Diego, Calif.*

An engineering paper on the C19K Charactron will include detailed discussions of its operating characteristics and applications to which this cathode-ray tube is being put as a large screen-target identification-display instrument in air surveillance systems. Its utility as a computer read-out device will also be discussed.

An analysis of the tube and its associated components has been made, which includes a discussion of the electron optical system, the beam-forming principle, and beam-deflection system. The discussion will be supplemented by diagrams, illustrations, and photographs which

explain the optical properties and behavior of the beam, and which also show actual displays of characters on the viewing screen of the tube.

11.5 "Versatility of Floated-Type Rate Integrating Gyroscopes in Systems Applications," by *J. W. Lower, Minneapolis-Honeywell Regulator Co., Minneapolis, Minn.*

Floated Rate Integrating Gyroscopes (HIG Units) have been gaining ever increasing acceptance in high-performance systems over the last five years. The reasons for this include the combination of extreme ruggedness with high sensitivity, a wide dynamic range of operation, fast response characteristics, and considerable versatility in servo-loop closures. These gyros can be used as displacement gyros, as platform-stabilization gyros, or as rate-measurement gyros. The combination of a precision torque motor with the gyro permits highly accurate precession characteristics to command inputs. Several gyro applications and the attendant systems' concepts are explored in this paper.

SESSION XII

TUES. 10:00 A.M.–12:30 P.M.

WALDORF-ASTORIA JADE ROOM

Broadcast Transmission Systems I TV Broadcasting

Chairman: RONALD J. ROCKWELL, *Crosley Broadcasting Corp., Cincinnati, Ohio*

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12.1 "Synchronization of Multiplex Systems for Recording Video Signals on Magnetic Tape," by *Donald E. Maxwell and William P. Bartley, G. E. Co., Syracuse, N. Y.*

The over-all and inter-channel timing-accuracy requirements for multiplex, magnetic-tape recording systems are discussed, with particular emphasis on the synchronization of time-division multiplex systems designed for use in the megacycle frequency range. The effects of tape flutter and skew on timing accuracy are considered, and methods for achieving good synchronization despite flutter and skew are described. The effects of inadequate synchronization on the signal output of the system are shown.

12.2 "Channel Response Requirements of Multiplex Systems for Recording Video Signals on Magnetic Tape," by *Benjamin G. Walker, General Electric Co., Syracuse, N. Y.*

Recently the time-division multiplex process has been employed as a means of recording a wide-band video signal on multiple-track magnetic tape. The desired frequency response in the individual channels is discussed, and a simple method for relating the sampling errors to the deviations from ideal response is developed. The effects of sampling errors on the output video signal are shown.

12.3 "Ferrite Heads for Magnetic Recording in the Megacycle Range," by *William R. Chynoweth, General Electric Co., Syracuse, N. Y.*

The performance, in a linear magnetic recording system, of ferrite heads having $\frac{1}{8}$ mil resolution, has been evaluated. The use of ferrite heads in a system recording in the megacycle range has been shown to be feasible, and a determination of the advantages and disadvantages of these heads for this use has been made.

12.4 "Attenuation Measurements on Short Line Samples," by *Louis E. Raburn, Crosley Division, AVCO Manufacturing Corp., Cincinnati, Ohio.*

A simple technique has been developed to measure the attenuation of relatively short samples of rigid coax lines. The general technique employed is to short both ends of the line, couple into one end through an impedance meter, and then measure the input impedance of the line as a resonant coaxial (TEM) cavity. In an example measurement the line sample is terminated at both ends with shorting plates having RMA flanges and center bullets.

The same technique could be used to calculate the attenuation of other transmission lines including the open-wire and waveguide types.

12.5 "A New Television Transmitting Antenna," by *R. W. Masters and C. J. Rauch, The Ohio State University, Columbus, Ohio.*

The structural simplicity, broadband impedance, and beam-pattern properties of traveling-wave antennas lead one to consider them for use in TV broadcasting. By locating the proper type of element more or less uniformly in a prescribed spiral along a transmission line, it is possible to obtain high-gain, null-less, pancake beams which are stabilized in direction over suitable vhf or uhf frequency ranges.

A practical embodiment utilizing probe-coupled, axial-slot radiators in the outer shell of a large diameter coaxial-transmission line is described, as well as measured characteristics of a representative (channel 7), high-frequency scale model, simple analysis and design equations.

12.6 "Spurious Emission Filters for High-Power TV Transmitters," by *William J. Judge, Jr., East Paterson, N. J.*

The basic theory and design of the transmission line constant-*K* low-pass section is evolved from a survey of papers on the subject. In addition, theory and design of transmission line low-pass *M*-derived sections (series and shunt) is discussed.

The composite filter is constructed, with specific application to television transmission service. Correlation of theory with practical design of high-power vhf and uhf filters is demonstrated. Associated mechanical problems are related. Charts showing the pass band and stop-band characteristics of (a) channel 2–6 25 kw, (b) channel 7–13 50 kw, and (c) channel 14–83 12 kw low-pass filters are supplied. Effect of insertion of the filter in the transmission system is outlined.

SESSION XIII

TUES. 10:00 A.M.–12:30 P.M.

WALDORF-ASTORIA SERT ROOM

Audio I General

Chairman: HARRY F. OLSON, *RCA Laboratories, Princeton, N. J.*

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13.1 "Electronically Controlled Audio Filters," by *L. O. Dolanský, Northeastern University, Boston, Mass.*

The use of filters whose cut-off characteristics are controllable by electronic means is often desirable in problems dealing with audio signals. Based on the recent work on fixed rc active filters by J. G. Linvill, variable active low-pass and high-pass filters have been developed using transistor negative impedance converters.

The design theory of such filters will be summarized and measured characteristics compared with those predicted theoretically. An application, in which the cut-off characteristics are controlled by the incoming audio signal for use in formant tracking, will be described and experimental results given.

13.2 "Distortion in Class B Transistor Amplifiers," by *Maurice V. Joyce, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

An experimental investigation has been made of distortion in Class B grounded-base and grounded-emitter amplifiers using junction transistors. While it is now known that a small amount of forward-emitter bias allows low distortion operation of such circuits, no detailed investigation of distortion as a function of bias has previously been published. This information, however, may be of great interest in very low-power drain circuits, where transistors would normally be used.

Data are also given on the dependence of distortion on source impedance, load impedance, and frequency. It is shown that distortion is practically independent of power output, depending rather on output current. Gain and input impedance are also discussed.

13.3 "Detection of Audio Power Spectrum Dispersion," by *H. S. Littleboy and J. Wiren, Northeastern University, Boston, Mass.*

This paper deals with the design and description of a "*V*" filter which is to be used to distinguish between power spectra having the same central frequency and average power but different dispersions. The power spectra that are discussed have a Gaussian distribution curve when plotted on a logarithmic frequency scale. The filter is adaptable for use in the analysis of speech and its use in this analysis will be described. The design of the filter including graphical means of establishing the filter characteristics from the dispersion of the given power spectra is included.

13.4 "Calibration of Test Records by *B*-Line Patterns," by *B. B. Bauer, Shure Brothers, Inc., Chicago, Ill.*

Test records are often calibrated by measuring the width of the optical pattern formed when a sharp beam of light is reflected from the modulated grooves. An error has been found in this measurement owing to diffraction of light at the edges of the pattern, which results in a fuzzy ending and a general enlargement of the pattern width, especially at high frequency. A new optical method has been devised for calibration of test records by the use of interference-line patterns. Two sets of interference lines have been identified: (a) Uniformly-spaced lines which are related to the recorded frequency, the angular velocity of the record, and the wavelength of light—these have been called the *A*-lines; (b) lines with variable spacing which are related to the amplitude of modulation and to the wavelength of light—these have been called the *B*-lines. *B*-line patterns may be readily related to the theoretical width of the optical pattern free of diffraction error, with the resultant improvement in the accuracy of test-record calibration.

13.5 "Design and Performance of a High-Frequency Electrostatic Speaker," by *Lloyd Bobb, R. B. Goldman, and R. W. Roop, Philco Corporation, Philadelphia, Pa.*

An electrostatic speaker has been developed which provides a quality of high-frequency reproduction not available with electro-magnetic tweeters. The diaphragm consists of a thin plastic film bearing an evaporated metallic layer. The membrane is stretched around a semi-cylindrical perforated electrode on which ridges are embossed to provide clearance. The response varies less than ±2 db in the frequency range between 8 and 16 kc. The azimuthal distribution pattern is excellent, owing to the cylindrical geometry, and is essentially independent of frequency in the same range. The second harmonic distortion inherent in this type of speaker is maintained at a low level. An indication of the quality of high-frequency reproduction is provided by oscillograms of the response to tone burst signals. The speaker is in quantity production and has been incorporated in several models of home reproduction instruments.

13.6 "Electronic Music Synthesizer," by *Harry F. Olson and Herbert Belar, RCA Laboratories, Princeton, N. J.*

The electronic music synthesizer is a machine that produces music from a coded record. The coded record is produced by a musician, musical engineer or composer with a fundamental understanding of the composition of sound. The electronic music synthesizer provides means for the production of a tone with any frequency, intensity, growth, duration, decay, portamento, timbre, vibrato, and variation. If these properties of a tone are specified, the tone can be completely described. The advantage of the electronic music synthesizer is that it can produce new and radical tone complexes for musical satisfaction and gratification. The new system does not displace the artist and musician of today. It does not take the place of talent combined with work. The electronic music synthesizer provides the musician, musical engineer and composer with a new musical tool with no inherent physical limitations.

SESSION XIV

TUES. 10:00 A.M.–12:00 Noon

WALDORF-ASTORIA GRAND BALLROOM

Information Theory I

Chairman: LOUIS A. DEROSA, *Federal Telecommunication Labs., Nutley, N. J.*

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14.1 "Coding For Noisy Channels," by *Peter Elias, Massachusetts Institute of Technology, Cambridge, Mass.*

A remarkable result of Shannon's "Mathematical Theory of Communication" is the fact that information can be transmitted at a definite rate over a noisy channel, with as low an error probability as is desired, by properly matching the message to the noisy channel. The matching process is called "coding." The problem of designing codes for particular noisy channels is central for many applications of the theory. This paper reviews the progress which has been made in coding, especially for the noisy binary channel. Some new results will be presented on the dependence of error probability and efficiency on coding delay and equipment complexity.

14.2 "The Rate of Approach to Ideal Coding," by *C. E. Shannon, Bell Telephone Labs., Murray Hill, N. J.*

Let C be the capacity of a noisy discrete channel without memory. Consider codes for the channel consisting of 2^{Hn} sequences each sequence being n symbols long. The basic coding theorem for a noisy channel states that by taking n sufficiently large it is possible to find codes such that both $\Delta = C - H$ and the probability of error after reception, P_E, are arbitrarily small. This paper is concerned with the problem of estimating the value of n necessary as a function of P_E and Δ. Both upper and lower bounds are found in fairly general cases. The upper and lower bounds for n approach each other percentage-wise as Δ and P_E approach zero, giving an asymptotic formula for n. Various special cases are investigated in detail, such as the symmetric binary channel and the flat gaussian channel.

14.3 "The Mathematics of Information Theory," by *Brockway McMillan, Bell Telephone Labs., Murray Hill, N. J.*

14.4 "Session Commentary," by *Robert M. Fano, Mass. Inst. of Tech., Cambridge, Mass.*

SESSION XV

TUES. 10:00 A.M.–12:30 P.M.

KINGSBRIDGE ARMORY MARCONI HALL

Instrumentation, Telemetry and Remote Control

Chairman: ROBERT L. SINK, *Consolidated Engineering Corp., Pasadena, Calif.*

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15.1 "Compound Modulation—Method of Recording Data on Magnetic Tape," by *George B. Newhouse, Consolidated Engineering Corp., Pasadena, Calif.*

This paper describes a new method of recording instrumentation data on magnetic tape. A combination of suppressed-carrier amplitude modulation and frequency modulation (compound modulation) minimizes the errors associated with direct recording and FM carrier recording. The theory and principles of operation of a typical CM channel are presented. It is shown that the system is compatible with FM and direct recording. Thus, all types of input transducers may be used. The errors for each system are compared and it is shown that CM has the following advantages: (1) small zero drift, (2) FM modulator center frequency not critical, (3) system nonlinearities have reduced effect, (4) less flutter error, and (5) good amplitude stability.

15.2 "Development of a Portable Magnetic Tape Recorder for Precision Data Recording," by *Glenn D. Maxwell, Consolidated Engineering Corp., Pasadena, Calif.*

Development of a portable, three-speed, 28-channel magnetic-tape recorder, primarily intended for application in an aircraft flight test instrumentation system having a system accuracy of 1 or 2 per cent, is described. The principal design requirements are compactness, reliability, low flutter, ruggedness, and versatility. About half of the contents relate to flutter. Definitions and methods of measurement are given. Results of a six-month flutter research program are presented. Equations show the effect of capstan torque variation and eccentricity on flutter. A rigid capstan drive is compared to a filtered drive. Results of environmental tests and design features are presented.

15.3 "A System for Precise Time-Storage and Expansion of Electrical Data," by *Clarence B. Stanley, Ampex Corp., Redwood City, Calif.*

A six-speed, four-channel magnetic-tape data recording and reproducing system for fixed installation is described. Solutions of design problems posed by rigid specifications on flutter, timing accuracy, operational and maintenance ease are discussed. The solution to the problem of precise timing accuracy in reproduction is of particular interest. A time displacement error of ±.25 milliseconds throughout the length of a 4,800-foot reel is achieved at the high-tape speeds, thus permitting precise recording and reproduction of time-based data.

15.4 "Automatic Oscillograph Readers," by *Louis L. Fisher and George L. Hatchett, J. B. Rea Co., Inc., Santa Monica, Calif.*

At the present time, considerable amounts of data are recorded on oscillographs. The analysis of these data is limited by the task of con-

verting the information into a form which can be handled by computing or tabulating devices.

This paper describes techniques which are useful to automatically read oscillograph records and provide a suitable output for subsequent processing. Originally developed for teledeltos-type paper, the principles evolved may be applied to other types of records with nonintersecting traces.

15.5 "Analysis of Data Recording Systems," by *Thomas L. Greenwood, Huntsville, Ala.*

Data recording is analyzed as to goals, methods, and techniques. System components are classified as functional elements. Definitions of system elements are suggested with a view to facilitate comparison between systems.

Ambiguities in current nomenclature of data handling systems are discussed, with a view toward standardization of terms.

Recording systems are treated as communication channels and their information-handling properties evaluated with the aid of current information theory.

The transducer is classified as an element but not discussed except in cases where the recording techniques are affected by the transducer type.

Proposals are made for a high-speed automatic digital-recording system using electronic counting and magnetic-tape recording techniques.

SESSION XVI

TUES. 10:00 A.M.–12:30 P.M.

KINGSBRIDGE ARMORY FARADAY HALL

Electron Devices I Tubes

Chairman: G. D. O'Neill, *Sylvania Electric Products, Inc., Bayside, N. Y.*

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16.1 "A Gas Discharge Noise Source," by *W. Honig and P. Parzen, Johns Hopkins University, Baltimore, Md.*

Previous studies have indicated the presence of large fluctuations in a gas discharge maintained by a magnetic field. An experimental study was made of the noise powers available from gas discharge tubes of Philips gauge geometry. Measurements indicate that noise powers as much as 70–80 db above KT per megacycle are available in certain frequency bands. Curves describing the operating conditions and frequency spectrum of noise power are given. It is believed that the large magnitudes of noise power are due to the presence of growing plasma waves which exist because of the magnetic field.

16.2 "Corrections to the Theory of the Grounded-Grid Triode," by *W. A. Harris, Radio Corp. of America, Harrison, N. J.*

In available literature on the theory of the grounded-grid triode, it has been assumed that the input conductance includes "the square-of-the-frequency" term caused by transit time in a grounded-cathode circuit and that feedback can be neglected. Even at moderate frequencies, the use of these assumptions leads to inconsistencies between calculated and realized performance.

This paper describes an exact method of treatment based on four-pole circuit equations. Expressions derived for the short-circuit admittances, using the Llewellyn-Peterson equations, are compared with measured data. Theoretical and measured values and curves are also given for power gain and for input admittance with tuned output loads.

16.3 "Development of a Large-Diameter Dumet Lead for Sealing to Soft Glass," by *D. L. Swartz and J. C. Turnbull, Radio Corp. of America, Lancaster, Pa.*

Dumet leads having diameters greater than 0.030 inch have long been desired in the power-tube industry for increased rf-current-carrying capacity. This paper describes a 0.060-inch, high-conductivity dumet lead which was developed by careful matching of radial and axial thermal-expansion coefficients. In this lead, relative copper volume is about ½ to ⅓ that found in conventional dumet leads. The core material is a nickel-iron alloy having an expansion coefficient of 90×10^{-7} in/in/°C. The use of this alloy provides a considerably better match to G0120 glass than that provided by conventional dumet leads. Relative merits of three possible techniques for producing the high-conductivity dumet leads are discussed.

16.4 "Novel Design Approach for Microwave Tubes," by *J. E. McLinden and D. Lichtman, Airborne Instruments Lab., Inc., Mineola, N. Y.*

An approach to electron tube construction is presented which is highly suited to tubes and components requiring establishment and maintenance of critical geometries in vacuum-sealed, heated assemblies. In this method, spacings are established by optical measurements and maintained by coherence with the assembly envelope. This approach should not only improve reliability of the component but reduce cost. The method is discussed as applied to the development of a planar microwave triode having spacing tolerances of ±0.0001 inch.

16.5 "Magnetron Operation at Very-Long Pulses," by *Markus Nowogrodzki, Amperex Electronic Corp., Hicksville, N. Y.*

The paper describes a phenomenological study of long-pulse performance in a 4J52-type magnetron employing a new type of emitter, the Philips' impregnated cathode. The basic cathode structure is described, a comparison of electrical performance with a standard tube at normal operating conditions is made, and the new tube's (6507) operation at 14-microsecond pulse conditions is analyzed. Design problems in the specialized test equipment are related, and pertinent tube data at long-pulse operation are described, such as cathode temperatures, power output, stability, life-test performance, et cetera. A tentative specification for long-pulse operation of the 6507 is included.

SESSION XVII

TUES. 2:30–5:00 P.M.

BELMONT-PLAZA MODERNE ROOM

Instrumentation II

Chairman: Edwin P. Felch, *Bell Telephone Labs., Inc., Whippany, N. J.*

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17.1 "A New Instrument for the Automatic Measurement of Transistor Noise Figure," by *D. D. Grieg and S. Moskowitz, Electronic Research Associates, Caldwell, N. J.*

The noise generated in a transistor is one of its basic characteristics, and the accurate measurement of this quantity is of considerable importance in the development, manufacture, and application of transistor devices.

This paper describes a new instrument which eliminates the drawbacks of the present manual procedure of transistor noise figure measurement. The instrument operates on the basic principle of synchronously connecting and disconnecting a calibrated noise generator to the input of the transistor or transistor-amplifier under test. By automatically comparing the internal transistor noise with that of the external noise source, a continuous direct reading of noise figure is obtained.

The paper describes the basic operating principles, gives details of a practical operating model, and presents comparative results with respect to standard manual methods.

17.2 "A Radio-Frequency Parameter Bridge for Junction Transistors," by *Anthony Hlavacek and Ge Yao Chu, Sylvania Electric Products, Inc. Ipswich, Mass.*

A radio-frequency bridge to measure the ohmic base resistance and the collector capacitance of junction transistors is described. The four arms of the bridge are essentially formed by a three-terminal rc network, and the collector-base circuit of the transistor under test. A sinusoidal radio-frequency signal is applied to the collector-base terminals while a null detector is placed between the emitter and a terminal of the rc network. The principle of operation and a brief description of the bridge are presented.

17.3 "A Versatile Transistor Tester for Measuring Open Circuit 'T' Parameters," by *R. P. Crow, Motorola, Inc., Chicago, Ill.*

Although "T" equivalent circuit parameters are conveniently used, and can be readily measured on point contact transistors, it has not been practical to make direct open-circuit measurements of these parameters on junction transistors because of the high-collector impedances. This paper describes a technique for using a "Q" multiplier for raising the effective collector-circuit impedance of a tester to values of over 200 megohms, permitting accurate

measurements of the open-circuit parameters. A description, with circuit diagrams, is given of a complete tester that measures 1,000 cps small-signal "*T*" equivalent circuit parameters and grounded-base and grounded-emitter current gains, input impedances with various loads, and power gains with various loads.

17.4 "A Transistorized Oscillograph," by *W. G. Reichert, Jr., Allen B. DuMont Labs., Inc., Clifton, N. J.*

The attractiveness of small size and low-power consumption of transistors led to the development of a transistorized, battery-operated, portable oscillograph. Utilizing only commercially available transistors, this unit has a frequency response of from 20 to 200,000 cps, a deflection factor of 0.2 volts per inch, and a sweep range of from 20,000 to 5 μsec. A description of the feedback amplifier, the multivibrator, and the turn-off and lock-out circuitry will be supplemented by a discussion of design problems peculiar to this application. High input impedance and output voltage swing requirements of the amplifier, and linearity restrictions on the sweep circuit will be covered.

17.5 "A 200 CPS to 5 MC Recording Equipment," by *Charles C. Comstock, Sig. Corps Engrg. Labs., Ft. Monmouth, N. J.*

The task of designing and constructing a magnetic-type recorder-reproducer to operate in the band 200 cps to 5 mc is divided into two portions. The mechanical system for achieving high relative head-to-tape motion with its attendant problems of obtaining proper head-to-tape contact and registration of reproduce heads on the recorded tracks, is one portion. The second portion is the electrical system to rearrange the signals which are to be recorded into such a form that the broad band may be recorded, and eventually rearranged at playback into the original frequencies.

SESSION XVIII

TUES. 2:30–5:00 P.M.

**WALDORF-ASTORIA
STARLIGHT ROOF**

**Engineering Management I
Panel Discussion: Operations Research—A Tool of Engineering Management**

Chairman: C. M. Jansky, Jr., *Jansky & Bailey, Inc., Washington, D. C.*

Panel Members: Sir Robert Watson-Watt, Airtron, Ltd., Toronto, Canada; Sherman Kingsbury, Arthur D. Little Company, Cambridge, Massachusetts; Dr. Ellis Johnson, Operations Research Office (Army), Johns Hopkins University, Baltimore, Md.; Leroy Brothers, Operations Analysis Group Hqts., U. S. Air Force; Martin Ernst, Operations Evaluation Group (Navy), MIT, Cambridge, Mass

SESSION XIX

TUES. 2:30–5:00 P.M.

**WALDORF-ASTORIA
ASTOR GALLERY**

**Aeronautical and Navigational Electronics II
Radar and Aircraft Landing Aids**

Chairman: Bernal M. Meador, *Trans World Airlines, Kansas City, Kans.*

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19.1 "Airport Surface Detection Equipment," by *J. E. Woodward, Airborne Instruments Lab., Inc., Mineola, N. Y.*

Airport surface detection equipment (ASDE) is a radar set used to observe surface traffic on an airport. Tests of an experimental model have demonstrated its usefulness but also demonstrated the desirability of making certain improvements in performance. These improvements have been achieved in a design program preceding the fabrication of an engineering model which will have the following:

Frequency	24,000 mc
Pulse Width	0.02 μsec
PRF	14,400 pps
Azimuthal Beamwidth	0.25 degree
Antenna Vertical Coverage	Horizon to −25 degrees
Antenna Polarization	Vertical or Circular
Scan Rate	60 rpm.

The technical aspects of this equipment and the special design problems will be discussed.

19.2 "A Marine Radar Identification System," by *Charles M. Tiffin, Federal Telecommunication Labs., Nutley, N. J.*

A marine radar identification system is described which operates on the principle of a transponder. The system enables a radar operator to distinguish the azimuth of a ship carrying the identification equipment from other ships in the vicinity.

The equipment is portable, self-powered, inexpensive, and simple to operate.

It is thought that this system, or variations thereof, will be useful in solving the radar identification problem as applied to marine or aeronautical navigation.

19.3 "Statistical Techniques for Analysis of ILS Flight-Test Data," by *Abe Tatz, Airborne Instruments Lab., Inc., Mineola, N. Y.*

Large amounts of flight-test data were obtained in an Instrument Landing System (ILS) evaluation program. These data were reduced by machine computation to provide quantitative measures of ILS performance. The measures of ILS performance are explained, the computing procedures are described, and typical results are presented to demonstrate the application of statistical techniques to a multitude of flight-test data.

19.4 "An Analysis of Angular Accuracy in Search Radar," by *Robert Bernstein, Columbia University, New York, N. Y.*

The problem of determining the theoretical limit of angular accuracy in search radar and the manner in which the controllable system parameters affect accuracy is of current interest in the design and operation of automatic track-while-scan systems, height-finding radars, and early warning detection and tracking systems.

The ultimate limitation to angular accuracy resides in the existence of receiver noise and target scintillation. Both these phenomena are random processes. The problem is formulated in a manner which brings the effect of these random processes into evidence, and it is concluded that the methods of parameteric estimation as developed in the field of mathematical statistics are most appropriate for its solution. Possession of a sufficient estimator of target azimuth would be most satisfactory because such an estimator would extract from the echoes all the information they contain concerning azimuth. It is proved, however, that a sufficient estimator of azimuth cannot exist for the case of search radar.

The maximum likelihood estimator is chosen as the best alternative and its distribution is found. A set of weighting functions are obtained for the purpose of maximum likelihood azimuth estimation; these functions are interpreted in terms of the information content of echoes from various parts of the antenna beam. Quantitative results are presented which give the standard deviation of the estimator as a function of beam axis rms signal-to-noise ratio and the number of pulses per beamwidth. This result is employed as a standard of performance against which the efficiency of more easily implemented azimuth estimators is judged.

19.5 "Radio Direction-Finding from the Standpoint of Sampling and Interpolation," by *Martin Masonson, Federal Telecommunication Labs., Nutley, N. J.*

This paper discusses systems ranging from the rudimental in radio guidance such as the Bellini-Tosi, to continuous scan systems, e.g. a VOR beacon, to versatile high-accuracy systems such as those presented by Earp and Godfrey. Concepts underlying and unifying an apparent variety of such systems, innovations on them, simple criteria and formulas for intrinsic repetitive errors are given and illustrated. From the standpoint of this paper these systems appear as interpolation computers which in practice are manifested in a diversity of instrumentations suiting given operational or production demands.

SESSION XX

TUES. 2:30–5:00 P.M.

**WALDORF-ASTORIA
JADE ROOM**

**Broadcast Transmission Systems II
Color Television**

Chairman: Robert E. Shelby, *National Broadcasting Co., Inc., New York, N. Y.*

20.1 "Proposed Controls for Electronic Masking in Color Television," by *W. L. Brewer, J. H. Ladd, and J. E. Pinney, Eastman Kodak Co., Rochester, N. Y.*

Electronic masking in the color television studio is accomplished by the use of cross-coupling networks. These networks yield three output signals, each of which is a linear function of the three input signals. The colors of the final reproduction depend upon the values of the coefficients in the linear equations relating output voltages to input voltages. A change in the value of *any* individual coefficient in these equations affects the hue, saturation, and brightness of *most* nonneutral colors. The effect of an incremental change in each coefficient is described for a given set of masking equations. Coefficient controls may be ganged to simplify the relationship between manual adjustments and visual effects. A block diagram is shown for a particularly suitable circuit. A means for ganging the controls is also described.

20.2 "Experimental Equipment for Recording and Reproducing Color Television Images on Black-and-White Film," by *William L. Hughes, Iowa State College, Ames, Iowa.*

This paper is concerned with the design and construction of certain pieces of experimental electronic and mechanical equipment which can be used for recording and reproducing color television images on black-and-white film. The design of this equipment is based on the black-and-white film system for color television proposed at the 1954 National I.R.E. convention. Certain changes from the original proposals are described.

Theoretical and economic advantages and disadvantages of such a system are discussed briefly, but the major portion of the paper is concerned with descriptions and operating characteristics of existing equipment.

20.3 "Cathode-Ray Vectorgraph," by *Frank Uzel, Jr., Allen B. Du-Mont Labs., Inc., Clifton, N. J.*

This paper will describe a new cathode-ray vectorgraph developed primarily to display NTSC color television chrominance information as processed by a vector decoder unit.

The specifications and performance characteristics of this unit will be presented. Of special interest are the stability and small relative phase shift of the identical X and Y amplifiers.

Although primarily intended for color television instrumentation, it is felt that this unit will have considerable general application where relative phase-shift measurements are necessary.

20.4 "Automatic Balance Control of Colorplexers in Color TV," by *J. R. Popkin-Clurman, Telechrome, Inc., Amityville, N. Y.*

A new colorplexer is described which has been designed to eliminate long and short time drifts of balance for chroma components, thus eliminating the necessity for the frequent manual adjustment of balance controls as in standard colorplexers.

An auxiliary correction servo unit is also described for application to any type of colorplexer.

An automatic balancing system takes advantage of the error components present in the color signal during the synchronizing interval. The encoder is then servoed to null the unbalance components. A memory system is incorporated so that interruptions in signals or switching transients will not interfere with the operation of the correction circuits.

The application of the principles described in the paper eliminates one of the main defects in present-day color broadcasting equipment.

20.5 "Television in Europe," by *Hubert A. S. Gibas, Uster, Switzerland.*

Three different television standards exist in Europe. These cause problems in the frequency allocation of television transmitters, in the construction of receivers for places where two or three television systems can be received, and in the international exchange of television programs. A European television network is in the state of development. There are, with a few exceptions, uhf links in use. The greatest progress of television we find in Great Britain, where, as early as the year 1937, a regular television service was inaugurated. In Europe, also, industrial television has found practical application.

SESSION XXI

TUES. 2:30–5:00 P.M.

WALDORF-ASTORIA
SERT ROOM

Audio II
Symposium: Music, High Fidelity, and the Listener

Chairman: LEO L. BERANEK, *Bolt, Beranek & Newman, Inc., Cambridge, Mass.*

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21.1 "Electronic Organ-Tone Radiation," by *Daniel W. Martin, The Baldwin Piano Co., Cincinnati, Ohio*

The acoustical effects achieved and the techniques used in the radiation of electronic organ tone are quite different from the conventional effects and practices of both public-address sound systems for auditoria, and "high-fidelity" music reproduction systems for the home. In this paper the fundamental acoustical design principles of organ tone chambers are reviewed. Specific examples of tone cabinet design are described, and organ installation principles are illustrated by examples of actual installation.

21.2 "The Role of Room Acoustics in Music Listening," by *John A. Kessler, Massachusetts Institute of Technology, Cambridge, Mass.*

As sound travels from a source to a listener or to a microphone, its spectral and temporal characteristics are altered by the room through which the sound travels. In the case of recorded music, at least two rooms usually contribute to the quality of the sounds which the listener hears. The effects of the room on the music may be beneficial or detrimental to listening enjoyment. The evaluation of these physical effects in terms of listener preference is a subject of continuing interest to musicians, architects, and engineers concerned with design of concert halls and with the recording and broadcasting of music.

21.3 "Environment-Fitness Considerations of High-Fidelity Audio Systems," by *R. D. Darrell, Stone Ridge, N. Y.*

Performance evaluations become progressively more difficult as the field of interest expands from isolated audio components to integrated systems and as rigorously objective measurements either become impractical or are superseded by wholly subjective aural judgments. Yet if the final evaluation of over-all sound qualities must be determined by aesthetic rather than engineering criteria, the latter still are significantly applicable to system-operation techniques and to the fitness of a specific system to a specific environment. The present paper calls for extended and intensified study of present-day operator-listeners' practical audio needs and psychological attitudes, and, anticipating the results of such study, suggests some possible engineering approaches to the better "matching" of home high-fidelity systems to their actual "habitats."

21.4 "Acoustic Requirements of a Sound System Determined by the Listener," by *Cyril M. Harris, Columbia University, New York, N. Y.*

The listener to a high-fidelity system requires a certain acoustical performance from his system. This paper discusses the acoustical factors which set the requirements the system loudspeaker output must be capable of meeting to be acceptable to the critical listener. Areas of research will be indicated which must be pursued in order to provide additional knowledge required to specify completely this problem.

21.5 "Man, a Somewhat Neglected Component of Hi-Fi Systems," by *Walter A. Rosenblith, Massachusetts Institute of Technology, Cambridge, Mass.*

In recent years much progress has been made in assessing the transmission efficiency of communication systems. In most situations that are of interest to the "hi-fi" enthusiast, it is however not possible to specify the message that is to be transmitted. Under these circumstances, one might suggest that the most realistic yardstick for the performance of hi-fi systems is man's discriminative ability. This talk will, in the main, deal with man's hearing. It will also be concerned with the question: "To what extent do laboratory experiments on pure tones predict human reactions in more general listening situations?"

SESSION XXII

TUES. 2:30–5:00 P.M.

KINGSBRIDGE ARMORY
MARCONI HALL

Telemetry and Remote Control III
Recent Telemetering Developments

Chairman: MARTIN V. KIEBERT, JR., *Convair, Pomona, Calif.*

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22.1 "A Multiple-Frequency Antenna-Coupling System," by *H. R. Sigler, Bendix Aviation Corp., North Hollywood, Calif.*

The most satisfactory solution to the problem of radiating several telemetering carriers from a vehicle is to radiate them from a single antenna, provided they can be coupled to it without excessive loss.

Two antenna couplers are suggested and described that will make up a system to couple as many as four transmitters to a single-phased antenna.

The first coupler is of the notch-filter variety and applies two signals to one antenna with a minimum insertion loss. The second coupler is based on the hybrid ring principle and will couple two signals to a phased-antenna array. When two of the former units are used with one of the latter, a system results which will combine four signals at one antenna.

The paper describes a new transmission line configuration which is used successfully in these two couplers and contributes to their compactness and resistance to environmental effects. The electrical characteristics and environmental reactions are also discussed.

22.2 "Germanium Photoconductor as Missile Spin-Counter in an All-Transistor FM/FM Telemeter Transmitter," by *C. M. Kortman, Bendix Aviation Corp., North Hollywood, Calif.*

A germanium photoconductor is used to frequency modulate a telemetering-subcarrier oscillator according to the intensity of the light falling upon the sensitive area. Intensities between complete darkness and bright sunshine can be measured and the difference between the light source and the shadow cast by the device, when it is between the source and the sensor, causes an appreciable frequency shift. In the missile application it is only necessary to count the cycles between bright light and shadow to determine the spin frequency. The germanium photoconductor is ideal for use with the all-transistor system since it requires no high-operating voltages or other special treatment. The frequency response is adequate for spin rates as high as several hundred revolutions per second.

22.3 "Linear Voltage Controlled Frequency Modulation of the Hartley Oscillator," by *W. F. Link, Bendix Aviation Corp., North Hollywood, Calif.*

A brief historical background of the evolution and use of the Hartley oscillator in FM/FM missile-telemetering systems is given.

The theoretical problems involved in obtaining large, linear-frequency deviations of an oscillator by voltage stimulus are discussed, particularly as affected by limitations imposed for use in missile telemeters. Practical solutions to this problem are given, especially as they have been achieved with the Hartley oscillator. Advantages accruing from use of the Hartley circuit over other possible oscillators are discussed. Finally a description is given of a commercially available voltage-controlled FM oscillator developed by the Pacific Division of Bendix Aviation Corporation.

22.4 "Application of Process Circuitry to Telemetering Components," by *L. A. G. Ter Veen, Bendix Aviation Corp., North Hollywood, Calif.*

The new arts of printed and etched circuitry have been applied to many items where high-quantity production and economy have been the most important factors. In telemetry components, neither of these two factors is of primary importance, reliability and compactness being worthy of greater consideration. This paper shows how process circuitry may be employed to enhance these factors.

The specialized design techniques which must be employed to adapt process circuitry to meet the needs of telemetering components is discussed in detail. Specific examples are given. A telemetering subcarrier oscillator employing an etched circuit, which is currently being produced in a quantity of 3,000 units, will be illustrated and described. Telemetering packages utilizing etched circuitry to achieve rugged, compact construction will also be discussed.

22.5 "Wide-Band AC Rate Networks," by *L. F. Lyons, Bendix Aviation Corp., North Hollywood, Calif.*

In servo system design, there frequently arises a need for lead or lag rate networks in the signal or feedback loop in order to achieve stability or modify the servo system frequency response for special applications. In systems where dc control and feedback voltages are employed, rate network design is comparatively simple. Many systems employ ac control and feedback because of its compatibility with synchro-position indicators and two-phase drive motors; however, ac differentiating or integrating networks present a more difficult design problem. Networks normally employed for this purpose are of the bridge *T* and parallel *T* null network, or lc resonant circuit variety. Both are frequency-drift sensitive, limited to high intelligence-frequency applications, and may have relatively high insertion loss.

This paper discusses ac lead and lag networks for applications at all intelligence frequencies. The networks presented are insensitive to variations in signal frequency. They also permit mixing of control and feedback signals in a common impedance which includes the rate network even in cases where one signal is ac and the other dc.

SESSION XXIII

TUES. 2:30–5:00 P.M.

KINGSBRIDGE ARMORY
FARADAY HALL

Electron Devices II
Microwave Tubes

Chairman: WELLESLEY J. DODDS, *RCA Tube Div., Harrison, N. J.*

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23.1 "Klystron Power Amplifiers for Long-Hop Microwave Relay," by *N. P. Hiestand, Varian Associates, Palo Alto, Calif.*

Recent interest in long-hop microwave relay communications has again emphasized the importance of the klystron as a means of obtaining very high power in the microwave region. This paper will review the use of klystrons as a communications instrument and discuss the latest developments in the field.

Particular emphasis will be placed on those features of the klystron amplifier that are important to the communications engineer, such as efficiency, gain, tunability, linearity, bandwidth, and operating parameters.

A family of tubes will be described which provide appreciable energy (several kilowatts) at most important communication frequencies from 350 mc to 8,000 mc.

23.2 "Wide-Band, High-Power Traveling-Wave Tubes at S-Band," by *S. F. Kaisel and W. L. Rorden, Stanford University, Stanford, Calif.*

This paper will describe three experimental traveling-wave tube power amplifiers in the 100 w to 1-kw power range at *S*-band. The T-230 and the T-231 are pulsed traveling-wave tubes which will give the order of 1-kw output at 30 db gain from 2.5–3.5 kmc. The T-231 employs a control grid in the convergent flow electron gun, to allow beam-keying with low-voltage pulses. The T-351 is a cw power amplifier which will give greater than 100 w of cw output at 20 db gain from 2–4 kmc. The electrical design considerations for such tubes and the mechanical design details will be described.

23.3 "A 1-KW Pulsed Traveling-Wave Tube Amplifier at X-Band," by *J. E. Nevins, S. F. Kaisel, and M. Chodorow, Stanford University, Stanford, Calif.*

This paper will describe the application of the contra-wound helix structure to a high-power, narrow-band pulsed traveling-wave tube amplifier for operation in the frequency range 8.5–9.6 kmc. Power output greater than 1 kw at 30 db gain has been obtained. The impedance of the contra-wound helix as predicted by Chodorow and Chu has been verified by cold perturbation methods and by beam measurements. Some comparisons are made with performance and dimensions that would be predicted for the single-wound helix. A novel method of construction for the cross-wound helix will be described.

23.4 "Noise Analysis of Traveling-Wave Tube Video Detector," *Glen Wade, Stanford University, Stanford, Calif.*

An analysis is made of the rectification characteristics and the noise characteristics of a traveling-wave tube modified for use as a video detector. The purpose of the analysis is to determine the "minimum detectable signal" as limited solely by noise.

The high velocity electrons in the velocity modulated beam at the output of a traveling-wave tube can be separated from the slower electrons by means of a velocity-sensitive deflecting field or by biasing the collector near cathode potential. The current thus separated can be developed into a rectified signal.

Using reasonable values for the tube parameters the calculated "minimum detectable signal" from such a device is approximately minus 80 dbm.

SESSION XXIV

TUES. 8:00–10:30 P.M.

WALDORF-ASTORIA STARLIGHT ROOF

Automatic Control II Trends in Automatization of Procedures and Processes in Business and Industry

Chairman: GORDON S. BROWN, *Massachusetts Institute of Technology, Cambridge, Mass.*

Panel Members: Gordon S. Brown, Massachusetts Institute of Technology, Cambridge, Mass.; Richard L. Meier, University of Chicago, Chicago, Illinois; W. R. G. Baker, General Electric Co., Syracuse, N. Y.; Low K. Lee, Stanford Research Institute, Stanford, Calif.; Roger W. Bolz, Penton Publishing Co., Cleveland, Ohio.

The growing use of the inter-connected techniques of feedback control, automatic assembly (automation), and electronic digital computation in industrial operations has led to an intense interest in the technical, managerial, economic, and social implications of automatic control. In this panel session, a chairman and speakers with a variety of backgrounds have been selected in order to obtain a comparison of viewpoints, as well as an over-all look at current trends toward automatization of procedures and processes in business and industry.

SESSION XXV

TUES. 8:00–10:30 P.M.

KINGSBRIDGE ARMORY MARCONI HALL

Audio III Seminar: Magnetic Recording for the Engineer

Chairman: SEMI J. BEGUN, *Clevite-Brush Dev. Co., Cleveland, Ohio*

25.1 "Magnetic Tape as a Recording Medium," by *Frank Radocy, Audio Devices, Inc., New York, N. Y.*

Magnetic tape effects on recorder characteristics are discussed. Value of new polyester base in overcoming temperature and humidity effects is studied. Oxide improvements and use of coating orientation have improved output and signal/noise ratio. New binder formulations have improved anti-friction characteristics even in high humidity, without impairing high-frequency response by cupping. Defect-free tape for computer, instrumentation, and video recording has developed from improved coating, slitting, winding, and quality control techniques, manufacturing not over one defect in 24,000 feet. Special dust-free packing protects in transportation and storage, but user must protect against dust-pickup on recorder.

25.2 "Recorder-Reproducer Design," by *Walter T. Selsted, Ampex Corp., Redwood City, Calif.*

This paper presents the various design perameters which must be considered in the design of a magnetic recorder intended particularly for radio industry use. Those design characteristics which must be weighed when designing a machine of high performance and reliability are studied in relationship to one another and in some considerable detail.

25.3 "Tape-Recording Applications," by *Marvin Camras, Armour Research Foundation, Chicago, Ill.*

Standard designs are flexible enough for most uses of tape recorders. Special machines have been devised for unusual applications, such as pronouncing dictionaries, length-measuring machines, time compressors, dc and square-wave recorders, artificial reverberation generators, musical instruments, memory devices, and video recorders. The construction and operation of typical devices are reviewed.

25.4 "Tape Life," by *W. S. Latham, Groton, Conn.*

This paper will be concerned with certain physical properties of magnetic tape and environmental factors which create tape-storage problems. Comparisons will be made of the effects of storage on instantaneous-lacquer disk recordings, with effects on magnetic-tape recordings stored for the same period of time under the same conditions of temperature and humidity. Actual measurements of coefficients of expansion and contraction of various types of magnetic tape due to temperature and humidity changes in storage vaults will be discussed.

Information will also be presented concerning the effects of continuous replay of recorded tapes upon the physical structure of current oxides and backing, and upon the information recorded thereon.

Essential requirements for the construction of a special tape and disk storage vault, at present under construction at the U. S. Navy Underwater Sound Laboratory, will also be presented.

25.5 "The Future of Magnetic Recording," by *John S. Boyers, National Company, Inc., Malden 48, Mass.*

Magnetic recording was invented somewhat over 50 years ago and has seen its greatest development in the past 10 or 15 years. Both systems and components are still in a terrific state of flux. Future developments in both areas are certain to be startling. In the area of components, possibilities for greater packing density, both linear and cubic are discussed along with transducers and other components. In the area of systems, application to automation controls and other non-audio as well as pure audio application are considered.

SESSION XXVI

WED. 10:00 A.M.–12:30 P.M.

BELMONT-PLAZA MODERNE ROOM

Ultrasonics I

Chairman: FRANK MASSA, *Massa Labs., Inc., Hingham, Mass.*

26.1 "Antenna-Type Transducers for Ultrasonic Flowmetering," by *R. C. Swengel, York, Pa.*

In the first application of the Bidirectional Ultrasonic Flowmeter, the measurement of hydraulic turbine discharge, it was necessary to design transducers that would distribute energy uniformly over a flow passage whose rectangular area was 390 square feet.

Difficulties in space limitations as well as obstruction to the flowing water led to the development of simple, antenna-type transducers. They consist of slender metal rods driven in compression by crystal or magnetostrictive drivers. Several methods for evaluating performance, both optical and electronic, are discussed, as well as specific data and the results of trials in a 42,500 horsepower hydroelectric turbine.

26.2 "Electrokinetic Hydrophones," by *Ernest Yeager, Western Reserve University, Cleveland, Ohio.*

At the interface between a solid and an ionic solution, a layer of ions of predominately one charge is absorbed on the surface of the solid with a diffuse layer of ions of predominately opposite charge adjacent in the solution. The periodic disturbance of this diffuse layer of ions by acoustical waves can produce several types of alternating potential effects which are significant in hydrophone applications. Such electrokinetic hydrophones offer special advantages at both infrasonic and ultrasonic frequencies in terms of their resistive impedance and their extremely small dimensions.

26.3 "Characteristics of Torsional Transducers," by *R. N. Thurston and Peter Andreatch, Bell Telephone Labs., Inc., Murray Hill, N. J.*

This paper is a collection of significant information about transducers for generating or detecting torsional oscillations.

The existing methods of making torsional transducers from barium titanate ceramic, quartz, and ADP are reviewed. Typical values are given for the electromechanical coupling coefficient. the "Q." and the voltage-torque

sensitivity when used as a torque-meter. Barium titanate ceramic transducers having coupling coefficients as high as 0.37 have been realized. Some practical methods of attaching these transducers to other materials such as glass and steel are described.

26.4 "Parameters Affecting the Q of Quartz Crystal Units," by *A. W. Warner, Bell Telephone Labs., Whippany, N. J.*

One of the principal electrical characteristics of a quartz crystal unit used as a high-frequency standard is its Q. Other factors being equal, the ability of a crystal unit to reduce the effects of circuit instabilities on the frequency of a crystal oscillator is directly proportional to its Q.

Studies have been made to find the relationship of Q to frequency, temperature, size, shape, surface preparation, and mode of vibration of AT-cut quartz plates, varying in size from 12 mm to 100 mm in diameter and 0.1 to 20 mc in frequency.

This has resulted in (1) crystal unit designs having a wide choice of crystal unit impedance in the frequency range of 1 to 10 mc, all units having a Q greater than 1×10^6, (2) a very practical primary frequency standard crystal unit design at 5 mc having a Q greater than 2×10^6, and (3) a particular design having a Q of 10×10^6 at room temperature. It is believed that Q's limited principally by the internal losses of the quartz itself have been achieved, and some data are given showing the relationship of these losses to frequency and temperature.

26.5 "The Frequency-Temperature Behavior of Piezoelectric Resonators Made of Natural and Synthetic Quartz," by *Rudolf Bechmann, The Brush Laboratories Co., Cleveland, Ohio.*

Resonators made of synthetic quartz show some differences in the frequency-temperature behavior compared with natural quartz. To evaluate the frequency-temperature behavior of resonators, the determination of the temperature coefficients of the measured frequency dependence up to the third order is necessary. These coefficients are functions of several variables, such as the orientation of the resonators and to some extent the ratio of dimensions as well as other factors. The representation of the frequency-temperature behavior of a resonator by the first-, second- and third-order temperature coefficients forms the basis for investigation into the fundamental constants, as the frequency-temperature behavior of a resonator depends on the temperature coefficients of the elastic constants and the coefficients of expansion. In synthetic quartz the constants are modified to some extent. These investigations also have practical application to problems of crystal manufacture. The synthetic quartz studied was grown under the U. S. Signal Corps sponsorship.

26.6 "Ultrasonics in the Decortication of Natural Fibers," by *Ethel R. Fleming, B. M. Harrison Labs., Newton Highlands, Mass.*

Experiments intended to evaluate the potential use of ultrasonic energy in the decortication of natural fibers are described. Practical limitations are discussed and possibilities for future developments along these lines are suggested. Particular emphasis is placed upon economic factors, especially insofar as these concern electronic generators. Essential requirements of high-frequency power supplies are examined.

SESSION XXVII

WED. 10:00 A.M.–12:30 P.M.

WALDORF-ASTORIA STARLIGHT ROOF

Electronic Computers I

Chairman: HARRY LARSON, *Ramo-Wooldridge Corp., Los Angeles, Calif.*

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27.1 "Experiments on a Three-Core Cell for High-Speed Memories," by *J. Raffel and S. Bradspies, MIT, Cambridge, Mass.*

A core memory ten times faster than the coincident-current memory (the fastest available arbitrary-access storage) now seems possible by the use of three cores per bit of information. None of these cores need be of the true square-loop type, since two are exclusively for switching, one exclusively for memory; it is merely necessary that the switch cores be saturable and that the memory core have remanence. Since selection is completely external to the memory cores, noise is extremely low. The speed of such a memory is limited only by the amount of current supplied to the cores. Single three-core cells have yielded cycle times below 0.5 microsecond, with access times between 0.1 and 0.2 microsecond.

27.2 "Bimag Circuits for Digital Data-Processing Systems," by *David Loev, William Miehle, John O. Paivinen, and Joseph Wylen, Burroughs Corp., Paoli, Pa.*

Some new techniques for using bimags in data-processing systems are disclosed. A brief summary of bimag characteristics is given, and three fundamental transfer loops are qualitatively described. Two of these transfer loops permit conditional transfer between bimags and thus permit logical operations upon isolated cores. A logical symbolism, useful both in drawings and system design, is offered.

Bimag circuits that are described include OR circuits and some novel techniques for realizing the AND, EXCLUSIVE-OR, and MATERIAL-EQUIVALENCE functions.

Some special-purpose magnetic shift registers are described, and techniques for synthesizing logical networks are indicated.

Comments on the experience gained in designing and operating a large bimag system are offered.

27.3 "A Transistor-Magnetic Core Circuit: A New Device Applied to Digital Computing Techniques," by *S. S. Guterman and W. M. Carey, Jr., Raytheon Manufacturing Co., Waltham, Mass.*

A circuit composed of a magnetic core with square-hysteresis-loop and of a single transistor, preferably of a junction type, forms a two-state memory cell.

In this circuit only the magnetic core remembers the state of the cell. The transistor serves as a driving or shifting element, as a buffer, and also supplies output power. Stand-by power is negligible. Diodes are not used in general.

From such cells are built all kinds of computing circuits, such as dynamic flip-flops, shift registers, counters—binary and decade, as well as logical structures.

These circuits are characterized by very low-power consumption, wide tolerances in component values, as well as in transistor parameters, and they have long life expectancy, especially if junction transistors are used.

Because the collector dissipation increases directly with frequency, the frequency of operation, as of this date, is limited by the transistor. With low-frequency low-power transistors, operation at 100 kc and even higher had proved to be satisfactory and quite reliable.

27.4 "A 'One-Turn' Magnetic Reading and Recording Head for Computer Use," by *D. F. Brower, Hughes Res. and Dev. Labs., Culver City, Calif.*

The design and characteristics of a "one-turn" magnetic reading and recording head are discussed. Among its desirable characteristics are high efficiency and ease of construction. An uncomplicated, compact, mechanical configuration simplifies shielding and the accurate positioning of gaps. Heads may be closely spaced while maintaining a low cross-talk ratio.

27.5 "Magnetic Selection Systems Using a Single Pyramid for Both Selective Writing and Reading in Large-Scale Electronic Computers," by *Amir H. Sepahban, Monroe Calculating Machine Co., Morris Plains, N. J.*

Selective writing of information on a chosen memory channel (magnetic drum channel) as well as selective reading from such a chosen channel in electronic computers is commonly performed by means of relay pyramids or diode or vacuum tube pyramids. This paper suggests the use of magnetic-core elements for this application in order to obtain a higher degree of reliability and longer life essential in large-scale electronic computers.

Description is given of the principle of operation of a 2-way all-magnetic 50-channel selection system developed as a prototype for a 3,500-channel magnetic drum storage unit.

Use of high permeability, nonrectangular hysteresis-loop-type cores allows noise-free gating of high-level writing currents as well as low-level reading voltages with a single magnetic pyramid. This departure from the conventional principle of switching cores with rectangular hysteresis-loop characteristics is essential for two-way operation of large-scale selection system.

Signal-to-noise ratios better than fifteen-to-one at a 100 kc signal repetition rate with channel selection speeds in the order of 100 microseconds are obtainable.

SESSION XXVIII

WED. 10:00 A.M.–12:30 P.M.

**WALDORF-ASTORIA
ASTOR GALLERY**

Microwave Theory and Techniques I Microwave Components

Chairman: H. F. Englemann, *Fed. Tel. Labs., Inc., Nutley, N. J.*

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28.1 "Wideband Waveguide Rotary Joint," by *Henry Schwiebert, Wheeler Labs., Great Neck, N. Y.*

A waveguide rotary joint has been developed for operation over a 20 per cent frequency bandwidth, matched within 1.2 db SWR (1.15 VSWR). It is based on a coaxial-line midsection, and handles the maximum pulse power consistent with single-mode propagation in such a line. At each end, it utilizes the ring type of junction with rectangular waveguide, previously developed for half this bandwidth. The improvements include a special compensation for the frequency variation of waveguide impedance and a special damping against incidental ring resonance in an uncoupled mode. Designs and models have been made for various bands in the range of 4 to 12 kmc.

28.2 "The Use of Modified Coaxial Structures for the Instrumentation of Components in Coaxial Line," by *Bernard Dwork, Harvard University, Cambridge, Mass., and Arthur A. Oliner, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

A number of components of novel construction employing modified coaxial structures have been designed for use in coaxial lines. These components are intended to supplement existing equipment so that a bench setup in coaxial line need no longer resort to transitions to rectangular waveguide for the necessary components. Among the components to be discussed are a variable attenuator, a hybrid tee, a directional coupler, and a novel type of double stub tuner. The modified coaxial structures employed are two- and three-plate parallel strip lines, and bifurcated coaxial lines. Emphasis in the design is placed on wide-band characteristics, small size, and simplicity of construction.

28.3 "High-Power Breakdown of Microwave Components," by *G. K. Hart and M. S. Tanenbaum, Sperry Gyroscope Co., Great Neck, N. Y.*

The main problems involved in the experimental determination of the peak-power capacity of waveguide components are discussed. An experimental approach based upon a statistical concept of breakdown is presented and the equipment required for this approach is described. Experimental evidence is given which supports both the statistical concept and the experimental approach. Data is presented on the power-carrying capacity of directional couplers, E- and H-plane tee junctions, E- and H-plane bends, and flange junctions.

28.4 "A Low Noise-Figure Microwave Crystal Diode," by *G. Messenger and C. T. McCoy, Philco Corp., Philadelphia, Pa.*

Microwave mixer noise figure can be uniquely calculated from the rectifier's equivalent circuit, which in turn can be uniquely determined from fundamental physics parameters.

The theoretical and practical relation of these parameters will be discussed in their application to the development of a germanium diode, with x-band noise figure as low as 4.5 db.

Though based on quantitative analyses, the presentation will confine itself to qualitative reasoning and graphical illustration.

Factors included in the discussion are the effects of spreading resistance, barrier capacity, power dissipation, noise temperature, surface treatment, etc.

28.5 "Tapered Velocity Couplers"—Part I, by *J. S. Cook.*

It is known that if two transmission lines with equal phase velocities are mutually coupled over some length, a signal introduced into one line will be completely transferred periodically back and forth between the lines.

This principle is employed in power-"splitting" devices where power is, for instance, equally divided between two lines. Half of the signal power is permitted to transfer from one line to the other and at this point the coupling is discontinued. Other devices utilizing this coupling principle provide complete transfer of a signal from one line to another. These devices, however, do not work so well when operation over a very large bandwidth is desired.

Certain considerations lead to the conclusion that if the relative phase velocities of two coupled transmission lines are caused to vary over their length, being equal at one end and widely divergent at the other, a signal introduced into one line will become equally divided between the lines. This scheme has only a second order frequency dependence and its useful bandwidth is a function of its physical length. A demonstration of a mechanical analog in the form of two coupled pendulums will be given to verify these conclusions.

Due to the relatively great physical length of this kind of coupler it will probably find most practical application at frequencies above 10,000 mc/s and/or in devices such as traveling-wave tubes with slow-wave structures.

28.5 "Tapered Velocity Couplers"—Part II, by *A. G. Fox.*

While a linear variation of the phase constant difference between a pair of coupled waveguides can produce a first order broadband transfer of power from one waveguide to another, the transfer characteristic will contain small ripples unless the waveguides are infinitely long. However, by simultaneously varying the coupling coefficient throughout the coupling region, the power transfer can be made substantially complete. The simultaneous variation of coupling coefficient and phase constant difference is based on a new principle of coupling between traveling waves here referred to as "normal mode tapering." The normal modes in such a tapered system are characterized by transverse field and energy distributions which change from point to point along the coupling region. Provided energy is introduced into only one of the normal modes, it can travel through the coupler with a desired change in the power distribution between two waveguides, and the distribution will not be sensitive to frequency changes. In contrast, the usual couplers employing matched phase constants achieve power transfer by means of the interference between two normal modes, and bandwidth will be limited by changes in this interference. A demonstration using the pendulum model will be given of the improvement in power transfer which can be achieved by simultaneous variation of coupling and phase constant difference.

SESSION XXIX

WED. 10:00 A.M.–12:30 P.M.

**WALDORF-ASTORIA
JADE ROOM**

Production Techniques—Electronic Equipment Assembly Methods

Chairman: Ralph R. Batcher, *Radio Electronic TV Mfrs.' Assoc., New York, N. Y.*

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29.1 "Electronic Design for a Digital Computer," by *R. J. O'Neill, Hughes Res. and Dev. Labs., Culver City, Calif.*

The cabinet is constructed of a top pan, bottom pan, and pillars separating these pans. The pillars are hollow, acting as supports and air ducts. Tiers are stacked between pillars and air vented through the tiers. Small chassis-holding electronic components plug into the tiers. Chassis are designed for preformed or postformed etched boards but presently are being used with flat boards until others are production tested. Machine wiring is all by taper pin connections. Cabinet doors open such that all wiring is accessible for service.

29.2 "A Flexible Automatic Component Assembly System," by *Ben Warriner and George W. Gamble, General Elec. Co., Ithaca, N. Y.*

This paper describes an automatic system for assembling lead-mounted components on printed circuit boards. It is being constructed under contract with the United States Army Signal Corps as an industrial preparedness measure. The component preparation, feed, placement, and testing functions are automatically controlled by a punch card programming system. Flexibility for job lot production has been stressed. This paper includes a discussion of the characteristics of the system, a description of the major units, and its overall operation.

29.3 "Principles of Circuit Packaging for Auto-Sembly," by *Sherman G. Bassler, Sig. Corps Engrg. Labs., Ft. Monmouth, N. J., and Myron Hinebaugh, P. R. Mallory and Co., Inc., Indianapolis, Ind.*

The printed-wiring technique of electronic circuit fabrication known as Auto-Sembly was introduced by the Signal Corps in 1949 and presented at the IRE Convention in New York in 1950. Acceptance and effective application by industry has confirmed the soundness of the principles involved.

The application of the system to more complex equipments, miniature as well as the larger size equipments, has shown the need for improvement in integration and circuit packaging methods.

To provide design information to meet this need has been the objective of a Signal Corps research and development contract with P. R. Mallory and Co., entitled "Development of Principles and Techniques of Integration and Packaging of Printed Circuit Assemblies," Contract DA-36-039 SC-42468.

This paper will present pertinent data developed from this contract and will show some of the aids evolved to assist the designer in selecting the proper circuit package and integration method to meet his design requirements. Four types of circuit packaging are presented with analyses of their efficiencies. Weight, volume, maintenance, and thermal efficiencies are considered. The methods devised for integrating the packaged sub-assemblies together with their integration efficiencies are shown.

The increasing demand for smaller, lighter, more rugged and efficient electronic gear, makes this new design information particularly useful for designers of new equipments.

29.4 "Standards for Automation," by *J. J. Graham, Radio Corp. of Amer., Camden, N. J.*

This is a paper discussing the need for and development of guide lines for the mechanization and eventual automatization of the assembly and construction of electronic equipment.

Development of standards for hole sizes, board sizes, component mounting and arrangement, equipment assembly, and manufacturing processes will be presented.

Comparison of problems and costs of automation processes with and without standards with special emphasis on machine design problems will also be treated.

The paper will close with a summation of the benefits derived from using standards, both from a manufacturing and a cost standpoint.

29.5 "Mechanization of Electronic Equipment," by *Frank B. Iles, Radio Corp. of Amer., Camden, N. J.*

This paper will present a discussion of the machine development program being followed by RCA's Production Department, with explanation of the over-all program and detailed description of machinery developed to date. A detailed description and production history of the following five pieces of equipment will also be furnished:

- 24-inch Shearing Machine with Automatic Control
- Programmed Punching Machine
- 12-inch Shearing Machine
- Two-Station Component Installing Machine
- Automatic Soldering Machine

Equipment usage will then be treated, with discussion of conveyorization and general plant arrangement for eventual automatization of the process.

29.6 "An Engineering Approach to Printed Circuitry and Automation," by *Rinaldo DeCola and George Harrigan, Admiral Corp., Chicago, Ill.*

This paper concerns the distribution of responsibility during the development of a printed circuit and automation program for the design and production of radio and television receivers using these techniques.

Also work over a period of time covering two and a half years is considered, certain specific problems and their solutions are discussed, equipment for use in the design and sample production of printed chassis, coils and other components is described.

Test and evaluation of field and environment conditions as well as field reaction and service problems are discussed.

A program of new component design, adaptation, quality control, and the need for further controls to insure more reliable flow of products through a high speed production system is considered.

As four groups, electrical engineering, mechanical engineering, production engineering and engineering quality control are involved in such a program. The co-ordination of these into an effective staff is covered.

While the sync, sound IF, video stage, and audio output tubes were first printed and mechanized, the adaptation of a universal IF system is described in detail.

SESSION XXX

WED. 10:00 A.M.–12:30 P.M.

WALDORF-ASTORIA SERT ROOM

Instrumentation and Nuclear Science

Chairman: URNER LIDDEL, *Bendix Aviation Corp., Detroit, Mich.*

30.1 "An Atomic Frequency Standard," by *Jerrold R. Zacharias, James G. Yates, and R. D. Haun, Jr., Massachusetts Institute of Technology, Cambridge, Mass.*

This research was supported by the Signal Corps; the Office of Scientific Research, Air Research and Development Command; and the Office of Naval Research.

The hyperfine structure resonance line of atomic Cesium at approximately 9192.632 mgc/sec has been used to stabilize the frequency of a microwave signal generator. This resonance frequency, describable as the Larmor precession frequency of the valence electron of atomic Cesium in the magnetic field of the atomic nucleus, is observed by the atomic beam magnetic resonance method as a resonance curve with a half-width of approximately 200 cycles/sec. The techniques necessary to make it into a practical device have become available only in the last few years in connection with researches on rare isotopes of the alkali atoms. These techniques will be described.

Preliminary results show that the short-time stability thus obtained (for times less than one second) is better than 1 part in 10^9 and that the average stability for long times will be considerably better than this. The beam tube has run satisfactorily for periods as long as 50 days indicating that a sealed-off tube with a life of several years should be quite possible.

30.2 "A Molecular Microwave Amplifier, Oscillator, and Frequency Standard," by *Charles H. Townes, Columbia University, New York, N. Y.*

A device has been developed which produces microwave amplification or oscillation as a result of interactions between electromagnetic fields and a beam of molecules in radiative states. The oscillation is extremely monochromatic and very stable in frequency because the frequency is primarily determined by the resonant frequency of isolated molecules. Experimental tests and theoretical analysis indicate that the device should provide a very constant and rugged frequency standard.

30.3 "Collision Reduced Doppler Effect. A Sodium Clock?," by *R. H. Dicke, Harvard University, Cambridge, Mass.*

It seems to be very important if not essential that an "atomic clock" employ an atomic or molecular resonance with an intrinsic frequency width which is very small. The one factor affecting this line width most difficult to modify is the Doppler effect resulting from atomic thermal velocities. The hyperfine transitions in the monatomic vapor of the alkali metals is characterized by the fact that the Doppler effect can be essentially eliminated in a very simple fashion. If the alkali metal vapor at a pressure of perhaps 10^{-6} mm of Hg is carried in a helium atmosphere at a pressure of perhaps .1 mm, the Doppler effect contribution to the line width is given by the time required by the alkali metal atom to diffuse a wavelength, rather than the time to move a wavelength as in the normal Doppler effect, an improvement of about a factor of 100. It is essential that the electronic spin orientation in the metal atom be unaffected by a collision with a helium atom. The technique of "optical pumping" for improving signal-to-noise will be discussed, also techniques for coherent pulse-induced radiation.

30.4 "Eddy-Current Bridge for Measurement of Skin Losses," by *Quentin A. Kerns, University of California, Berkeley, Calif.*

Search for a suitable measurement technique for skin losses led to the development of the present instrument, which seems to be one member of a family of similar devices.

The eddy-current bridge has been used to measure the conductivity of materials from silver to graphite at frequencies ranging from 1 kc to 40 mc. To obtain a reading, the sample is placed against the instrument head and two adjustments are made, corresponding to resistive and reactive balancing of a bridge, which may be calibrated to read resistivity in ohm-cm or ohms square at a given frequency.

A wide frequency range is desirable in the instrument. In particular, a plot of loss vs frequency for a sample can be used to measure plating thickness, or to detect the presence of a lossy layer on a base metal of higher conductivity. Such a lossy layer, found in copper-clad steel, was easily removed by pickling.

30.5 "Modifications to the Hutchinson-Scarrott Pulse Height Analyzer to Obtain a Coded Decimal Presentation and a Decimal Print-out," by *J. L. McKibben, J. D. Gallagher, and H. J. Lang, Los Alamos Scientific Lab., Los Alamos, N. M.*

The modifications to be described improved the memory stability to the point where the analyzer would give dependable operation for a period of at least one month with no adjust-

ments of the memory or adder during that time. The modifications to the memory section were made to improve the reliability and to permit an easy conversion to storage in a coded decimal system. The storage in a coded decimal system resulted in two advantages, an easily read display and the availability of data in a form suitable for use with an automatic print-out machine. The code was made such that the very simple binary adder could be retained. To further improve reliability, a very stable fused quartz delay line was chosen instead of a mercury delay line or a magnetostriction line.

SESSION XXXI

WED. 10:00 A.M.–12:00 Noon

WALDORF-ASTORIA GRAND BALLROOM

Symposium on Spurious Radiation

Chairman: Ralph Bown, *Bell Telephone Laboratories, Murray Hill, N. J.*

Panel: W. R. G. Baker, General Electric Co., Syracuse, N. Y.; A. V. Loughren, Hazeltine Corp., Great Neck, N. Y.; G. E. Sterling, formerly FCC Commissioner; Ernst Weber, Polytechnic Inst. of Brooklyn, Brooklyn, N. Y.; E. M. Webster, FCC Commissioner.

SESSION XXXII

WED. 10:00 A.M.–12:30 P.M.

KINGSBRIDGE ARMORY MARCONI HALL

Circuit Theory II General Theory

Chairman: Samuel J. Mason, *Massachusetts Institute of Technology, Cambridge, Mass.*

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32.1 "A Generalization of Foster's and Cauer's Theorems," by *F. M. Reza, MIT, Cambridge, Mass.*

The following lemma is proved by the potential analog method.

If $P(s)$ is a Hurwitz polynomial, then $d^nP(s)/ds^n$ is also a Hurwitz polynomial. Using this lemma as a starting point, the well-known Foster and Cauer's theorems are then generalized. The generalization in the latter case requires a more elaborate proof than in the first case. It is shown that if $Z(s)=P(s)/Q(s)$ represents the driving-point impedance of an RL or RC passive network, then

$$[d^nP(s)/ds^n] \div [d^nQ(s)/ds^n]$$

also will represent an RL or RC driving-point impedance.

A general discussion on the generation of positive real functions presenting the driving-point impedances of two-element kind networks is undertaken. In this direction the following general theorem is proved: If

$$Z(s) = \frac{a_0 + a_1s + \cdots + a_ns^n}{b_0 + b_1s + \cdots + b_ns^n}$$

represents the driving-point impedance of a two-element kind one-terminal pair dissipative network, and $P(s)$ is a Hurwitz polynomial with negative real distinct roots, then:

$$Zk(s) = \frac{a_0 + a_1P' + \cdots \cdot a_nP^{(n)}}{b_0 + b_1P' + \cdots + b_nP^{(n)}}$$

will also represent the driving-point impedance of a two-element kind one-terminal pair dissipative network.

32.2 "On the Separability of Laplace Transform Variable and its Applications in Carrier Systems," by *Sheldon S. L. Chang, New York University, New York, N. Y.*

Carrier systems are analyzed by separating the Laplace transform variable into a data component and a carrier component. Independent phasors are introduced giving rise to independent complex planes for the data transform variable and carrier transform variable. Data transfer function linking directly the demodulated signal to the modulating signal is derived for arbitrary carrier network and is applicable to both sinusoidal and transient inputs.

The method is applied to carrier servomechanisms. The equivalent time constants for a carrier network is derived under condition with phase and frequency errors.

For dual channel carrier servos, doubly complex input and response functions are introduced. Together with the complex transfer function, "cross-talk" can be treated as a single servo problem instead of as that of two interlinked servo-systems, with much simplification in the analysis.

32.3 "A New Approach to the Approximation Problem," by *Walter L. Baker, Ordnance Research Laboratory, Pennsylvania State University, State College, Pa.*

The signals encountered in industrial control processes and other control systems is often some arbitrary function which is known only in the form of graphical or tabulated data. It is desirable for design purposes to obtain a mathematical expression for the signal function. A method is presented for obtaining by a function of the form $e^{-t}(a_nt^n+a_{n-1}t^{n-1}+\cdots a_0)$, a desired fit to a large group of signal functions. It is assumed that the signal is a real function of time which:

1. is bounded and single valued for $0 \leq t < \infty$
2. must approach zero as $t \to \infty$
3. is zero for all values of $t<0$
4. has only points of ordinary discontinuity.

The method of approximation which is presented, provides greater flexibility and rapidity in obtaining a fit where the difference between the given function and the approximating function is to be minimized at any desired points. One can also estimate initially how much work is justified in obtaining a desired fit and arrange the detailed procedure accordingly.

32.4 "A New Series Representation for Correlation Functions," by *W. M. Kaufman and J. B. Woodford, Carnegie Institute of Technology, Pittsburgh, Pa.*

A special series representation of the autocorrelation function of a statistically described signal is presented which can adequately describe in a few terms most functions likely to be encountered. Use of this approximation will allow many synthesis problems to be solved analytically instead of numerically or graphically. A generalization of this method is suggested for the fourth and higher moment correlation functions. Examples are given of the use of this method in the design of optimum linear servomechanisms and filters involving a statistically described input signal.

32.5 "Theory of Low-Frequency Oscillators Employing Point-Contact Transistors," by *B. J. Dasher, D. L. Finn, and T. N. Lowry, Georgia Institute of Technology, Atlanta, Ga.*

This paper describes a theory for low-frequency oscillators, employing point-contact transistors, novel features of which include the use of precalculated design curves for determining the amplitude of oscillations, and the direct calculation of frequency stability. The analysis is based on a Fourier-series representation in which the nonlinear negative resistance characteristic is approximated by straight-line segments. Instead of assuming the characteristic to be nearly linear, the oscillation is assumed to be nearly sinusoidal. With the help of design curves computed for idealized typical characteristics it is possible to predict with reasonably good accuracy both the amplitude stability and the frequency stability of an oscillator with respect to known variations of transistor parameters.

The application of the theory to design problems is illustrated by example, and experimental confirmation of the results is presented.

SESSION XXXIII

WED. 10:00 A.M.–12:30 P.M.

KINGSBRIDGE ARMORY FARADAY HALL

Antennas and Propagation III Panel Discussion: Extended Range VHF and UHF Propagation

Chairman: Jerome B. Wiesner, *Massachusetts Institute of Technology, Cambridge, Mass.*

Panel Members: Kenneth Bullington, Bell Tel. Labs., New York, N. Y.; William E. Gordon, Cornell University, Ithaca, N. Y.; Oswald Villard, Stanford University, Stanford, Calif.; Dana Bailey, Nat. Bureau of Standards, Washington, D. C.; Walter Morrow, Massachusetts Institute of Technology, Cambridge, Mass.

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Discussion of new techniques, experimental data and current research in long-range vhf and uhf propagation.

SESSION XXXIV

WED. 2:30–5:00 P.M.

BELMONT PLAZA MODERNE ROOM

Ulstrasonics II

Chairman: Oskar E. Mattiat, *Clevite-Brush Dev. Co., Cleveland, Ohio*

34.1 "Nondestructive Testers by Means of Ultrasonics," by *Bertram M. Harrison, B. M. Harrison Labs., Newton Highlands, Mass.*

A general examination is made of the techniques currently being employed in nondestructive testing by the use of ultrasonic and sonic vibrations. Limitations economic and practical are discussed. Suggestions of possibilities involving new and different techniques are advanced and practical limitations of strictly theoretical approaches discussed.

34.2 "Ultrasonic Echo-Ranging for Tissue Diagnostic Studies," by *John M. Reid and John J. Wild, Medico-Technological Research Dept., St. Barnabas Hospital, Minneapolis, Minn.*

An ultrasonic beam is used to scan tissues at a carrier frequency of 15 megacycles, a pulse length of 1 microsecond, and a repetition rate of 1 kilocycle. A thickness-resonant, air-backed, *X*-cut quartz crystal is used as a transducer. Echoes returned by tissue are displayed, using radar techniques, to form a picture of the irregularities. Several transducer transport mechanisms are described. Empirical studies using biological techniques have been made of reflection patterns from different parts of the body. The equipment has been in clinical use for the diagnosis of human breast lesions, including cancer. Pictures characteristic of various types of breast lesions are shown.

34.3 "Techniques Used in the Ultrasonic Visualization of Soft Tissue Structures of the Body," by *Douglas H. Howry, University of Colorado Medical Center, Denver, Col.*

Sound echo-ranging techniques used in materials testing do not lend themselves to the study of the interior construction of the body, since the high acoustic absorption plus the anatomic complexity require a low-frequency signal while maintaining an azimuth and range-resolving power of 32 objects to the inch over a 40 db scale.

The range problem has been satisfactorily solved by three separate methods: (1) Either by coupling the crystal driver to water by making it into an asymmetric compound oscillator, or by a special acoustic impedance transformer, which produces an efficient wide-band transfer of power with a resultant short pulse. (2) By double reverse pulsing which causes the echo to phase-cancel itself after a predetermined number of oscillations. (3) By a new system of phase-modulation echo-ranging which makes possible a resolution of structures separated by ½ wave length.

The azimuth problem has been partially solved by beam focusing, and by a system of multiple position scanning which accurately interchanges range information for azimuth resolution. A computer circuit is being developed which will further improve azimuth resolution. Examples of two and three dimensional pictures will be shown, and technical and physical limitations of the system described.

34.4 "Technical Aspects of the Cavitron Ultrasonic Process in Dentistry," by *Lewis Balamuth, Cavitron Equipment Corp., Long Island City, N. Y.*

The Cavitron ultrasonic dental machine will be described with reference to its ultrasonic engineering features. Recent successes in clinical applications on patients have pointed to ultrasonic dentistry as one of the more significant advances in the growing field of ultrasonic technology.

The ultrasonic machining process will be discussed in the light of its meaning for operating on living tooth structure. The differences between the action of conventional rotary dental instruments and the Cavitron dental tool will be described in detail. Evidence will be presented to show that the Cavitron method of cutting tooth structure takes place at very low speeds and consequently with only the slightest pressure during the excavation. This is in direct contrast with the trend of rotary instrument research where recent work has been at high and higher rotary speeds.

34.5 "Application of Ultrasonics to Clinical Dentistry," by *Alvin E. Strock, Peter Bent Brigham Hospital, Boston, Mass.*

The use of ultrasonic vibrations to activate tools for carrying out dental procedures is being evaluated. It is proving of practical value in many branches of dentistry.

Using the Cavitron process, many intricate excavations in tooth structure can be made. The sensation of the ultrasonic drill to the patient is quite different and distinctly less disagreeable than conventional tools.

It can also be used for filing nerve canals in nonvital teeth. Conventional techniques have often resulted in freezing and fracturing of rotating files inside the tooth root. Files threaded into the end of the transducer are very effective and eliminate these difficulties.

Tools can also be made for condensing various types of filling materials. The possibilities for cutting bone and disintegrating tartar have great promise and will be reported.

34.6 "Ultrasonic Destruction of Erythrocytes," by *Eugene Ackerman and David B. Lombard, Pennsylvania State University, State College, Pa.*

Equipment has been constructed to observe the breakdown of mammalian erythrocytes in cavitating acoustic fields from 50 kc to 1 mc. These fields are produced in three ceramic bowls which can easily be used at frequencies either side of their resonances. The sound pressures are measured with calibrated ceramic probe hydrophones, and the particle velocities with electrokinetic hydrophones. The erythrocyte concentration is measured directly in the exposure tube in the sound field by means of an optical densitometer built into the equipment. The output is recorded on an Esterline Angus milliameter.

Preliminary measurements on the frequency dependence of the breakdown rate of the erythrocytes will be presented.

SESSION XXXV

WED. 2:30–5:00 P.M.

WALDORF-ASTORIA STARLIGHT ROOF

Electronic Computers II Symposium: Design of Machines to Simulate the Behavior of the Human Brain

Chairman: HOWARD E. TOMPKINS, *Burroughs Res. Center, Paoli, Pa.*

Panel Members: Anthony G. Oettinger, Harvard Computation Lab., Cambridge, Mass.; Warren S. McCulloch, Massachusetts Institute of Technology, Cambridge, Mass.; Nathaniel Rochester, International Business Machines Corp., Poughkeepsie, N. Y.; Otto H. Schmitt, U. of Minnesota, Minneapolis, Minn.

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The most versatile machines—electronic digital computers—can perform only a fraction of the tasks commonplace to the human brain.

How can we go about designing machines of greater versatility and adaptability, whose performance resembles more closely the higher behavior patterns of the human brain?

The possibility of such an approach is important both in engineering and in medicine, and the points of view of both of these disciplines will be discussed by the panel.

SESSION XXXVI

WED. 2:30–5:00 P.M.

WALDORF-ASTORIA ASTOR GALLERY

Microwave Theory and Techniques II Microwave Techniques

Chairman: W. L. PRITCHARD, *Raytheon Mfg. Co., Newton, Mass.*

36.1 "A Broadband Electronic Doppler Simulator," by *Gershon J. Wheeler and John Reed, Raytheon Mfg. Co., Newton, Mass.*

Instead of using a wheel or other mechanical device in this Doppler simulator, a single side-band signal is obtained electronically with no moving parts.

The incoming signal is split in a hybrid tee. Both halves are amplitude modulated by audio signals in quadrature and then recombined in quadrature. As a result one side-band is completely suppressed while the other is transmitted with little attenuation. Carrier suppression may be accomplished by using balanced modulators.

Extreme bandwidth is obtained by making the side arms of the hybrid of waveguides of different dimensions, so chosen that the difference in electrical length remains constant over a broad band.

36.2 "A Contribution to Microwave Measurements," by *F. J. Tischer, Huntsville, Ala.*

This paper is concerned with the subject of field distribution in the vicinity of a special helix line structure with regard to the existence of waves traveling with reduced velocity along the line element. The existence of such waves

makes practical the application of the line element to a standing wave detector at frequencies less than 500 mc.

Problems discussed include instruments which show the standing wave pattern of the field strength in a transmission line on the screen of a cathode ray tube. Prototypes are shown in photographs.

The investigation of the field distribution in waveguides, with both progressing and reflected waves, shows the applicability of a probe rotating around its axis for the standing wave detection.

36.3 "Measurement of Electromagnetic Parameters by the Use of Spheres Placed Near a Wall in a Resonant Cavity," by *W. K. Saunders, Diamond Ordnance Fuze Labs., Washington, D. C.*

A standard method for the measurement of electromagnetic parameters at microwave frequencies is through the study of the frequency shifts caused by the introduction of a sample into a resonant cavity. This method requires a knowledge of the field within the sample in terms of the unperturbed field. For rods and plates, this is easily obtained, but for the more mechanically convenient spheres, a problem arises as the sample is allowed to approach a wall of the cavity. Now the free-space dielectric sphere solution is no longer valid, and one requires a solution of Laplace's equation for a sphere in the presence of a plate. The author has not been able to find this solution in the literature. A method is given which indicates the structure of the solution and the variation of the cavity frequency perturbation as the sample approaches the wall.

36.4 "Impedance Measurement Through a Discontinuity in a Transmission Line," by *R. Mittra, University of Toronto, Toronto, Canada.*

The paper describes a method for calibrating a slotted line when measuring impedance through a discontinuity, e.g. an adapter, a coupling element between two guides, or a four-terminal network in series with the transmission line.

The method consists of performing standing wave measurements through the discontinuity with three specified reactance terminations. For the lossless junctions, three calibration constants are obtained directly from the measurement data. The evaluation of constants for a lossy structure involves certain simple graphical constructions on a Smith chart.

Typical application of the method in the fields of impedance measurements and matching, with examples illustrating the procedure, and certain advantages of the method over those already existing in the field are presented.

36.5 "Measurement of Small Complex Reflection Coefficient," by *Howard Scharfman, Raytheon Mfg. Co., Bedford, Mass.*

An accurate method of measuring reflection coefficients of magnitudes in the range .003 to .1 with good phase information is presented. The length of time required for a measurement is negligibly longer than for conventional slotted line techniques, and the measurement accuracy is considerably better. The system employs a ferrite circulator to separate incident and reflected powers at the unknown load. A microwave bridge is used to measure the relative phase and amplitude of the reflected power from the unknown load as compared to a reference short circuit.

The limitations of the method, means for extending the range of accurate measurement, and examples of successful application in matching problems are discussed.

SESSION XXXVII

WED. 2:30–5:00 P.M.

WALDORF-ASTORIA JADE ROOM

Quality Control and Reliability Studies of Electronic Tubes and Systems

Chairman: MARCUS A. ACHESON, *Sylvania Elec. Prod., Inc., Kew Gardens, N. Y.*

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37.1 "Prediction of Missile Reliability," by *M. J. Kirby and H. R. Powell, Sperry Gyroscope Co., Great Neck, N. Y.*

This paper describes a method of testing missile elements to failure and analysis of the results. The "ultimate strength" of an element is plotted as the probability of failure vs the duration of exposure to the test environment. This is a convenient numerical indication of the inherent reliability of a design or a production lot, if the simulation of the use-environment in the tests is realistic. Comparison of the curves of "probability of first failure" with the "probability of second failure" indicates what limits the reliability and the degree of improvement to be expected from a remedy. Also, classification of failures according to cause as "design," "workmanship," "tubes," etc. shows where corrective emphasis should lie and whether the proportion of manufacturing errors is decreasing or increasing. This makes the method useful for monitoring the quality of production. A sufficient number of missile components has been tested to failure to demonstrate the feasibility of the method, and its effectiveness in providing a basis for recommending remedial changes.

37.2 "Detection of Intermittent Circuit Faults," by *Sidney Wald, Glenn L. Martin Co., Baltimore, Md.*

The detection of intermittent faults in electronic equipment is an art in which the chief tools have been experience, ingenuity, intuition, and patience.

A new test method is described herein which is designed to ferret out potential trouble areas in electronic apparatus before the faults therein cause malfunction of the equipment (present operational checks and voltage and resistance analyses show only present faults, not potential ones), and without resorting to damaging extremes such as excessive heat, cold, moisture, shock, or overvoltage.

In its present state of development, this new technique applies to RF and IF circuits which normally handle continuous modulated signals.

The method consists of the following steps: (1) A large amplitude unmodulated carrier of frequency suitable to the circuit under test is applied to the circuit input; (2) the output is brought to a detector; (3) the detector output is amplified and monitored with a loudspeaker, test oscilloscope, and voltmeter. Refer to the block diagram (Figure 2); (4) the circuit under test is explored by gently tapping each component; (5) any perturbation of circuit constants causes a modulation of the carrier and results in an output. The circuit under test becomes extremely microphonic; that is, detachable electrical outputs are produced by extremely small mechanical vibrations or shocks. Such outputs are not produced in the absence of the carrier signal in most circuits.

37.3 "Statistics of Electronic System Failures," by *J. H. Parsons, K. L. Wong, and A. S. Yeiser, Hughes Aircraft Co., Culver City, Calif.*

Electronic systems are studied with respect to the probability of an operation of specified length being failure free. How this probability depends upon the maintenance policy is described. Under many types of maintenance, failures tend toward a pseudo-random distribution quite irrespective of the components' failure characteristics. Thus, the effectiveness of a proposed program of scheduled component replacement is difficult to predict from system failure statistics, but can be deduced if the coefficients of variation of the mean lives of the components are known. A simplified mathematical model of the failure behavior of a large scale electronic system is presented. Failure and error data from a moderately large digital computer are found in agreement with this model.

37.4 "New Reliable Voltage Reference Tubes for Severe Environmental Conditions," by *Earl J. Handly, Raytheon Mfg. Co., Newton, Mass.*

The variation in characteristics of some new glow-discharge voltage reference tubes are discussed. The paper includes material on the ionization and operating voltages for both short- and long-term operation. The effect of several factors such as vibration, shock, and high ambient temperatures are discussed. Values of initial voltage drift and voltage repeatability are given. The results are particularly applicable where the VR tubes are used as a source of reference voltage.

37.5 "Guided Missile Reliability and Electronic Production Techniques," by *Alfred R. Gray, Glenn L. Martin Co., Baltimore, Md.*

Reliability and maintenance problems peculiar to airborne electronic equipment are discussed. Emphasized are the fantastic demands made upon electronic component-part reliability by the vibration and temperature of guided missiles. Comment is made regarding the influence of component-part changes upon production techniques, as parts manufacturers strive to meet the missile reliability challenge. A general description is given (with slides) of various contemporary manual production methods, and of several concepts for electronic automation. An attempt is made to view these techniques and concepts objectively, and to evaluate them in terms of their ability to cope with present component-part changes, and in terms of trends for the future. A note of caution is sounded concerning blind acceptance of any one of these schemes as the *complete* solution

for present automatic production. But, at the same time, a note of confidence is voiced that out of these concepts of automation and semi-automatic production, will soon come a philosophy which will successfully marry automation techniques to the reliability requirements of even the intercontinental guided missile.

SESSION XXXVIII

WED. 2:30–5:00 P.M.

WALDORF-ASTORIA SERT ROOM

Broadcast and Television Receivers

Chairman: EDWIN B. HASSLER, *Warwick Mfg. Corp., Chicago, Ill.*

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38.1 "A Developmental Pocket-Size Broadcast Receiver Employing Transistors," by *D. D. Holmes, T. O. Stanley, and L. A. Freedman, RCA Labs., Princeton, N. J.*

This paper describes a pocket-size developmental AM broadcast receiver which utilizes eight junction transistors. Its performance is comparable to that of conventional personal receivers. Emphasis has been given to developments which contribute to stability with respect to temperature, battery voltage, and variations among transistors. The superheterodyne circuit employed uses a single-transistor frequency converter to perform the functions of both mixer and oscillator. Refined detector and automatic-gain-control circuits and an audio amplifier embodying further development of the principle of complementary symmetry are incorporated. Reduction in physical size and battery requirement, as compared to conventional receivers, is substantial.

The circuits are described in detail and certain aspects of components and of physical arrangement, which contribute to the small size, are discussed. Detailed performance data are also included.

38.2 "Progress in Ferrite Components for Television and Radio Receivers," by *H. M. Schlicke, Allen-Bradley Co., Milwaukee, Wis.*

The reported improvements in ferrite components for television and radio receivers are based on better material, or on novel circuits, or on the utilization of normally unwanted properties. An example for each kind is given.

A new high efficiency ferrite (W-02) perceptibly out-performs all others for color television flyback transformers. The half-cycle criterion is used to determine the optimum perameters of ferrites for this application.

A simple small sized omnidirectional ferrite antenna circuit provides a possibility for inside automobile antennas and miniaturized aircraft antennas.

Ferrite beads acting essentially as resistors for UHF only are incorporated in discoidal feed-through capacitors as integrated filter units.

38.3 "What Price Horizontal Linearity?," by *Monte Burgett and John Tossberg, Philco Corp., Philadelphia, Pa.*

This paper will be a study of the factors affecting the linearity of horizontal deflection. The types of nonlinearity are classified according to what part of the raster is affected and according to what circuit elements cause the nonlinearity. Statements are made of what changes in component parts and what additional parts are required to improve linearity. Estimates are made of the increased power drain and cost involved in these changes. The experimental techniques which were used in the study are explained.

38.4 "A Compatible High Definition Monochrome Television System," by *Pierre M. G. Toulon, and Francis T. Thompson, Westinghouse Research Labs., East Pittsburgh, Pa.*

A system for doubling the maximum number of horizontal and vertical picture elements of a televised scene is described. This system operates with conventional sweeps and is compatible with monochrome receivers.

The additional picture definition is obtained by making more efficient use of the available bandwidth. Redundancies between a given picture element in successive fields and between adjacent elements in a given field are measured at the transmitter, and this information is transmitted to the receiver in a compatible manner. This information is used to determine whether the receiver will display conventional large-size picture elements on a short-persistence phosphor or small-size picture elements on a long-persistence phosphor.

38.5 "Determination of the Optimum Demodulation Angles in Color Receivers," by *Stephen K. Altes, General Elec. Co., Syracuse, N. Y.*

A pair of demodulators can extract all the information from the color subcarrier signal. The only restriction on the output signal of a demodulator is that it should be zero for $R=G=B$, which represents reference white.

With the aid of a graph, the expression for the output signal for a certain demodulation angle can be obtained the most conveniently. The graphical method for accomplishing the inverse operation is also possible.

Detection circuits using angles other than the conventional $R-Y$ and $B-Y$ angles have advantages sometimes.

Several circuits incorporating these principles are analyzed.

38.6 "A Color Projection Receiver," by *W. F. Bailey and R. P. Burr, Hazeltine Corp., Little Neck, N. Y.*

This paper describes the results of an investigation of the use of projection displays for color television receivers. The display described consists of three small picture tubes producing red, green, and blue images and the Schmidt optical systems which project the three images onto a viewing screen. A general discussion is given of the technical problems encountered in the investigation and the means by which these problems were solved. One of the major problems—maintenance of registry—is considered in some detail. Some comparisons are made of the performance of the projection display with respect to other color display devices. Circuit description is limited to those circuits which differ from normal practice in color television receivers.

SESSION XXXIX

WED. 2:30–5:00 P.M.

KINGSBRIDGE ARMORY MARCONI HALL

Circuit Theory III Filters and Lines

Chairman: WILLIAM H. HUGGINS, *Johns Hopkins University, Baltimore, Md.*

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39.1 "A Method of Rational Function Approximation for Network Synthesis," by *N. De Claris, Massachusetts Institute of Technology, Cambridge, Mass.*

A general method is developed for finding functions of a single complex variable s, which approximates an assigned network characteristic, within the special class of functions realizable as networks of linear lumped parameter circuits. The method is based upon an interpolation technique with a series of general rational functions on the unit circle of the z-plane. A number of transformations that map the interval of interest of the s-plane into the unit circle of the z-plane are discussed.

A great advantage of the method presented in this paper is that it allows preassigning, if so desired, the pole location of the desired rational function anywhere in the left half of the s-plane. In addition to this freedom, it also makes it possible to approximate on some contour of the complex plane the function itself rather than its modulus. Indeed in some applications of electric circuits, one is interested in both magnitude and phase characteristics of a network in specified frequency interval.

Following a former mathematical treatment, procedures are outlined for three cases of approximation in frequency and time domain.

39.2 "Input Capacitance of Maximally-Flat Filters," by *John L. Stewart, University of Michigan, Ann Arbor, Mich.*

Low-pass lossless ladder filters, operating between different source and load resistances and adjusted to have an n-pole maximally-flat transfer function where n is the total number of inductors and capacitors in the filter, usually are built with a shunt capacitor as the element nearest the source and, if n is odd, also have a shunt capacitor adjacent to the load. These capacitances are evaluated for arbitrary n and source-to-load-resistance ratio using a method obtained from the Darlington synthesis procedure. Several formulas and useful curves are given. The relationship between the leading capacitance of the filter and the Bode limit is also demonstrated. Finally, it is shown that no mismatch is permitted if the maximum bandwidth-times-transmission-coefficient product is desired.

39.3 "Application of Time Series to the Calculation of the Transient Response of Bandpass Systems," by *C. J. Peters, Gulf Res. and Dev. Co., Pittsburgh, Pa., and J. B. Woodford, Carnegie Institute of Technology, Pittsburgh, Pa.*

A method is developed for determining numerically the response of a bandpass system to an arbitrary input, by use of a time series. The method is well suited to machine computation, and is an extension of low-pass time-series methods of others. Extreme complexity of the input, which can be either a time function or a time function modulating a carrier, is no disadvantage in using this method. As an example, the response of a spectrum analyzer is calculated for an input consisting of two sinusoidal signals of three different frequency separations for three different IF amplifier selectivities.

39.4 "Maximizing the Bandpass Ratio in Impedance Transforming Filters," by *D. H. Geipel, North American Aviation, Inc., Downey, Calif., and R. L. Bright, Westinghouse Elec. Corp., Pittsburgh, Pa.*

A lumped-element, four-terminal, impedance-transforming network is synthesized by applying a necessary condition to the impedance determinant of a system of mesh equations. The requirement for single band-pass operation is applied, and the relationships between bandwidth ratio and impedance transformation ratio are determined. Formulas for the element values of different types of three mesh networks as functions of the impedance transformation ratio are developed. Also, experimental results are compared to theoretical response curves.

39.5 "Miniaturized High-Impedance Magnetic-Core Delay Line," by *H. W. Katz and R. E. Schultz, General Elec. Co., Syracuse, N. Y.*

The high permeability and high resistivity of the ferrites have long attracted their use to the miniaturization of high-frequency magnetic components. One among many such components is the distributed constant delay line. Since the time delay per unit length equals $\sqrt{LC}$, large increases in the per unit length inductance by means of high permeability cores would decrease the total length of line required for a given time delay. But the simple introduction of ferrites into standard lines does not seem to lead to satisfactory results.

This paper will present an analysis of magnetic-core delay lines which will lead to the source of the poor high-frequency response which is normally obtained. The analysis also indicated what changes in the line configuration must be made to extend the frequencies over which the time delay remains constant.

Ferrite delay lines designed according to the procedures outlined in the analysis have been constructed and tested. Successful lines having a time delay of 0.2 μs/in. and a characteristic impedance of 6,000 ohms have been made. The useful frequency range has been extended to over 4 mc. In addition it has been found possible to vary the time delay electrically without serious pulse distortion. The paper will conclude with a discussion of the experimental results to date and an indication of future development.

SESSION XL

WED. 2:30–5:00 P.M.

KINGSBRIDGE ARMORY
FARADAY HALL

Antennas and Propagation IV
Propagation

Chairman: KENNETH A. NORTON, *National Bureau of Standards, Boulder, Colo.*

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40.1 "Airborne Measurement of Effective Ground Conductivity at Low Frequency in Alaska," by *Glenn M. Stanley and T. Neil, University of Alaska, College, Alaska.*

Airborne field strength measurements have been made over large sections of the Territory of Alaska in the frequency range of 203 to 407 kc. Theoretical field strengths are computed after the methods presented by K. A. Norton for both a plane and spherical single-layered earth and the theoretical values are compared with experimental data to obtain the effective ground conductivity. A discussion of the method of analysis is presented, and a map of the effective conductivity of wide regions of the Territory is included. The effective conductivity is shown to be between 0.5 and 5 mmho/m over most sections of Alaska.

40.2 "Atmospheric Attenuation of Microwave Radiation," by *Gene R. Marner, Collins Radio Co., Cedar Rapids, Ia.*

Solar radiation at wavelengths of 8.7 mm and 1.9 cm has been monitored for several months at the Collins Feather Ridge Observatory with two radio telescopes. The Collins Radio Sextant is utilized to position the two radio telescopes so that they continuously receive maximum solar power independent of atmospheric refraction. The decrease in received power as the atmospheric path length increases at sunset is used to calculate the atmospheric attenuation constant. A statistical study of the daily values of attenuation constants is presented, and the effect of several weather factors upon attenuation is discussed.

40.3 "Back-Scattering from the Sea Surface," by *Martin Katzin, Naval Research Lab., Washington, D. C.*

The available experimental data on back-scattering from the sea surface ("sea clutter") are described, and discussed with reference to the physical processes involved. It is shown that the scattering elements of the surface are illuminated by a combination of direct and indirect fields, the latter being "reflected" (forward-scattered) by surface elements closer to the radar. At low angles, these interfere destructively, giving rise to the so-called "critical angle," polarization dependence, "spikyness," and steeper frequency dependence. A theory is developed in which elemental scatterers are taken to be small patches or "facets" of sea surface. It is shown to give characteristics in general agreement with available experimental data.

40.4 "Measurements of Correlation, Height Gain, and Path Antenna Gain at 1,046 Megacycles on Spaced Antennas Far Beyond the Radio Horizon," by *A. F. Barghausen, M. T. Decker, and L. J. Maloney, National Bureau of Standards, Boulder, Colo.*

Studies made and reported by the National Bureau of Standards in 1953 on the correlation of 1,046 mc radio fields on spaced antennas at distances far beyond the radio horizon are supplemented by more recent measurements. Data taken from three vertically spaced dipole antennas for longer periods of time allow correlations of instantaneous fields to be determined for various fading rate conditions. Measurements were made on the National Bureau of Standards Cheyenne Mountain transmission path in Colorado and Kansas at a frequency of 1,046 mc, the receiving location being at an angular distance of 27 milliradians (for standard atmosphere), i.e. approximately 150 miles beyond the transmitter radio horizon. In addition to the correlation of instantaneous fields, the correlation of hourly median fields and the diurnal variation of the received signal are shown. Simultaneous measurements from a parabolic reflector antenna are used to indicate loss in gain relative to expected free-space gain of a large aperture array.

40.5 "An Airborne Radar and Wave Propagation Laboratory," by *David L. Ringwalt, Naval Research Lab., Washington, D. C.*

The Naval Research Laboratory has instrumented an R5D aircraft (Douglas DC-4) as an airborne radar and wave propagation laboratory. The installation includes four radar systems (with frequencies ranging roughly from 1 to 10 kmc), associated radar recording and calibration apparatus, and an extensive meteorological instrumentation for measurement of related atmospheric conditions. All radars operate synchronously from a centralized control. The antennas, along with transmitter-receiver systems, allow transmission of vertical, horizontal, or circular polarization, with simultaneous reception of transmitted polarization and its orthogonal component.

This aircraft is intended for investigations in two general fields, back-scattering (radar) and wave propagation. In the radar field, it may be used for the determination of the back-scattering properties of individual targets or of area-extensive targets such as ground, sea, or precipitation. In the propagation field, it may be used as one terminal of a one-way transmission link, such as in investigations of radio ducts and atmospheric scattering.

SESSION XLI

THUR. 10:00 A.M.–12:30 P.M.

BELMONT-PLAZA
MODERNE ROOM

Medical Electronics I
Panel Discussion

Chairman: OTTO H. SCHMITT, *University of Minnesota, Minneapolis, Minn.*

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Panel Members: Stanley Briller, Bellevue Medical Center, New York, N. Y.; Coleman C. Johnston, Lexington, Ky.; Joseph Moldaver, Institute of Physical Medicine and Rehabilitation, New York, N. Y.; J. W. Buchta, University of Minnesota, Minneapolis, Minn.; Britton Chance, University of Pennsylvania, Philadelphia, Pa.; Hector F. Skifter, Airborne Instruments Lab., Inc., Mineola, N. Y.; Saul R. Gilford, Colson Corp., Elyria, Ohio.

A panel discussion on medical electronics has been planned with the hope of being able to answer several questions which have been raised concerning this rapidly developing field of science.

The term, medical electronics, will be defined and its scope will be described. Doctors of medicine and electronic engineers will be brought together for a lively and extemporaneous discussion which will demonstrate good teamwork and a breakdown of the supposed language barrier. A timely and interesting problem will be presented to the members of the panel. An analysis of this problem will be made and likely solutions will be suggested. The value of the assumed results will be emphasized.

SESSION XLII

THUR. 10:00 A.M.–12:30 P.M.

WALDORF-ASTORIA STARLIGHT ROOF

Information Theory II

Chairman: Thomas P. Cheatham, *Melpar, Inc., Watertown, Mass.*

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42.1 "Time-Varying Filters for Nonstationary Signals on a Finite Interval," by *Arnold H. Koschmann, Univ. of Minn., Minneapolis, Minn.*

The concepts of signal space are applied to the problem of determining the optimum linear filter for a nonstationary signal which is known on the time interval (O, T). The output of the filter at time T can be expressed as the weighted sum of the Fourier coefficients of the input defined with respect to a suitable set of orthogonal functions. For a fixed value of T this sum can be interpreted as the convolution of the input with the impulse response of a fixed network. As T is varied, a set of impulse responses is obtained. Relationship between this set of impulse responses and impulse response of a time-varying linear system is shown.

42.2 "Analysis of Linear Systems with Randomly Varying Inputs and Parameters," by *A. Rosenbloom and J. Heilfron, The Ramo-Wooldridge Corp., Los Angeles, Calif., and D. L. Trautman, Hughes Aircraft Co., Culver City, Calif.*

This paper begins with a brief discussion of the historical progression of analyses of linear system behavior from fixed parameters and deterministic (completely known) signals to randomly varying parameters and signals. Major attention is then focussed on some of the newer techniques and latest results, due to the authors and others, applicable to (a) random inputs (not necessarily Gaussian) to fixed systems, and (b) deterministic inputs to randomly varying systems. In both of these cases the outputs are random processes which can be described by moments or more completely by probability distributions. Both modes of description are discussed.

42.3 "Detection of Coherent and Noncoherent Pulsed Signals," by *R. F. Drenick, S. Gartenhaus, and P. Nesbeda, RCA-Victor Division, Camden, N. J.*

The use of statistical methods in the theory of detection of pulse signals in noise has been very fruitful. Certain rules for optimum detection have been derived by various authors for the video as well as the IF portion of the receiver. This paper discusses the optimum detection procedure in the IF portion of the receiver. It extends previous work in that the cases of coherent and noncoherent detection of a signal of unknown frequency are considered. The problems are formulated statistically as a composite hypothesis problem and solved by Wald's decision function theory. The optimum (minimax) decision rules for the above cases are derived and illustrative mechanizations which suggest themselves in each case are described.

42.4 "The Linear Filtering of Sampled Data," by *Gene Franklin, Columbia University, New York, N. Y.*

The following is an application of least-squares filtering theory to situations where a time-stationary random message and additive noise are sampled before being filtered. Such a situation might possibly occur with a pulsed radar or data link or in a system where an instrument measures only samples. The transfer function of the filter obtained from this application of the theory is usually a product of two terms, one of them rational in s and the other rational in e^{sT}, where s is the complex frequency and T is the sampling period. The first term is obviously a network of lumped R, L, and C elements while the second term has the form of a linear program on a digital computer.

SESSION XLIII

THUR. 10:00 A.M.–12:30 P.M.

WALDORF-ASTORIA ASTOR GALLERY

Electron Devices III Cathode-Ray Type Tubes

Chairman: Russell R. Law, *CBS-Hytron, Danvers, Mass.*

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43.1 "A Time-Sampling and Amplitude-Quantizing Tube," by R. P. *Stone, C. W. Mueller, and W. M. Webster, RCA Labs., Princeton, N. J.*

The possibility of a saving of bandwidth—or of transmitting additional information in a given bandwidth—by means of amplitude quantizing and time sampling has recently been realized.

Beam-deflection type tubes were successfully built and tested, which perform these functions. They will change a continuous signal into a quantized signal having six discrete amplitude levels, and simultaneously sample as often as ten million times per second. A residue signal is also generated.

Two types of output structure have been used, both of which permit the external adjustment of the output amplitude levels. The tube operates with an anode voltage of 300 volts. The stability of operation is such that after initial setup no critical operating conditions or adjustments are involved.

43.2 "Cathode-Ray Tube with Single Step Intensifier," by *Jenny E. Rosenthal, A. B. Du Mont Labs., Passaic, N. J.*

A study is presented of a single-step intensifier for cathode-ray tubes. Various designs are investigated to test the premise that the electron lens constituted by the second anode and the intensifier should be a thin lens type to minimize distortions of the barrel or pincushion type. Field plots are given showing the fields corresponding to various electrode shapes, such as the truncated cone, suggested by standard electron lenses. Data are given on experimental tubes and compared with results obtained from field plots.

43.3 "An Electrostatic Shutter Image Converter Tube," by *B. R. Linden and P. A. Snell, A. B. Du Mont Labs., Inc., Passaic, N. J.*

The simple electrostatic image converter proposed by Schagen, *et al.* has been modified by the insertion of a mesh between cathode and anode to allow pulsing the tube. Design equations have been derived which will allow construction of the tube for any desired set of geometrical and electrical parameters. Tubes have been constructed which may be pulsed on and off with a 165 volt pulse for 10,000 volts between cathode and anode. With modifications in the geometrical parameters it is possible to construct a tube which may be pulsed on and off with as low as a 10 volt pulse for 10,000 volts between cathode and anode. The resolution of the displays obtained so far on this type of image converter is about 700 television lines over a 4-inch diameter circle. The photocathode sensitivity lies in the range of 30 to 60 μA/lumen to a light source at a color temperature of 2,870 degrees K.

43.4 "A New High Efficiency Parallax Mask Color Tube," by *M. E. Amdursky, R. G. Pohl, and C. S. Szegho, Rauland Corp., Chicago, Ill.*

Electron transmission through parallax masks of current tri-color tubes is no more than 12 per cent with a consequent low picture brightness. A new tube is described employing a parallax mask at a potential intermediate to that of the screen and a collector mesh maintained at anode potential. Not only is the brightness of such a tube increased 3–4 fold by virtue of enlarged mesh holes and the ensuing post-deflection focusing, but secondary emission from the mask, which would dilute color, is minimized. In contrast to other post-accelerating tubes, mask holes and fluorescent screen dots are uniformly spaced over the entire area. Nineteen-inch round and twenty-four-inch rectangular tubes incorporating the new principle have been built.

43.5 "The Tricolor Vidicon—an Experimental Camera Tube for Color Television," by *P. K. Weimer, S. Gray, H. Borkan, S. A. Ochs, and*

H. C. Thompson, RCA Labs., Princeton, N. J.

A television camera tube capable of generating three simultaneous color signals is now being developed. The experimental Tricolor Vidicon to be described is comparable in size to a standard monochrome Image Orthicon and has separate output terminals for each color channel. Color filter strips with associated conducting signal strips are built into a photoconductive target. All signal strips corresponding to the same primary color are connected to a common output terminal. Special preamplifiers have been developed to obtain independent color signals in the presence of the high interstrip capacity of the target. Color fidelity independent of the scanning process is obtained with a single electron beam.

SESSION XLIV

THUR. 10:00 A.M.–12:30 P.M.

WALDORF-ASTORIA JADE ROOM

Component Part I Electro-Magnetic Devices

Chairman: REUBEN LEE, *Westinghouse Elec. Corp., Baltimore, Md.*

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44.1 "Blocking Oscillator Transformer Design," by *P. R. Gillette, K. W. Henderson and K. Oshima, Stanford Research Institute, Stanford, Calif.*

An important step in the design of a blocking oscillator circuit is the determination of the coupling capacitance and transformer characteristics necessary to produce the desired output pulse. The nonlinear behavior of the tube characteristics cannot be taken into account by most methods that have been suggested for use in analyzing blocking oscillator circuits, but a step-by-step method has been developed which gives reasonably accurate predictions of output pulse shape. This method has been used in the calculation of the coupling capacitance and transformer turns ratio, open-circuit inductance, leakage inductance, and distributed capacitance required to produce an output pulse of optimum shape, maximum amplitude, and specified rise time and duration.

44.2 "Improvements in Pulse-Switching Reactor Design," by *R. A. Mathias and E. M. Williams, Carnegie Institute of Technology, Pittsburgh, Pa.*

Design equations are derived for quantity of core material and distribution of core material, copper in windings, and capacitors in Melville-line pulse-generating systems for microsecond and fractional microsecond service. Experimental tests described yield excellent agreement with predicted results. Examples of designs for particular high-power pulsing applications are also described.

44.3 "Magnetostriction Resonators as Circuit Elements," by *R. T. Adams, Federal Telecommunication Labs., Nutley, N. J.*

A compact magnetostriction rod resonator is described, applicable to a number of tone generator and filter circuits for carrier telephone and telemetering systems. The unit has properties of a single resonant circuit with independent input and output coupling. Frequency stability is ± 2 ppm per degree C, with typical Q's of 3,000 to 4,000 at 24 kc.

In addition to conventional applications as a "single-frequency" filter or as a frequency determining element, analagous to a quartz crystal, by suitable adjustment of magnetic bias, the resonator can be made to respond to sum and difference frequencies in the input and/or output circuit. A single rod can thus be used as a filter-modulator, as a modulator-filter, or as a complete heterodyne-tuned filter (modulator-filter-demodulator) tunable by means of an external local oscillator signal supplied to input and output. Efficient frequency doubling can also be performed by the element.

44.4 "Wide-Range Electrically Tunable Oscillators," by *John L. Stewart and Kermit S. Watkins, University of Michigan, Ann Arbor, Mich.*

This paper describes several recently explored low-level electronically tunable oscillators which illustrate techniques using conventional components and tubes for achieving rapid electrical tuning over relatively wide frequency ranges. The mechanism by which voltage (or current) control of frequency is achieved for each of the oscillators is described with particular attention being given to a most promising technique defined as parallel-network tuning. Typical circuitry as well as experimental curves relating output voltage and frequency as functions of the control voltage for the tunable oscillators are presented and practical limitations regarding gain, bandwidth, and component requirements are discussed.

44.5 "Fluorochemical Liquid and Gases as Transformer Design Parameters," by *L. F. Kilham, Jr. and R. R. Ursch, Raytheon Mfg. Co., Waltham, Mass.*

Liquid and gaseous fluorochemicals were applied with considerable success to the design of high temperature (Class H) transformers. We found these materials approximately equal in dielectric strength and power factor to 100 cstks, silicone oil, and "self-healing." Being chemically inert two or more of these liquids may be mixed to give a "tailor-made" dielectric-coolant. It is shown that such a liquid may be designed for maximum heat transfer for a specified condition of temperature and pressure.

Techniques using these dielectric-coolants are detailed and were used to redesign three specific Class "A" R.M. Co. transformers. Savings in size and weight by the redesigns are considerable. A magnetron filament transformer was reduced greater than 4 to 1 by volume.

SESSION XLV

THUR. 10:00 A.M.–12:30 P.M.

WALDORF-ASTORIA SERT ROOM

Nuclear Science I

Chairman: JEROME D. LUNTZ, *McGraw-Hill Publishing Co., New York, N. Y.*

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45.1 "A Study of a Variable Frequency Cyclotron Resonant System," by *M. R. Donaldson, R. E. Worsham, and N. F. Ziegler, Oak Ridge National Lab., Oak Ridge, Tenn.*

To accelerate several different particles in a cyclotron, the ratio of the frequency of the electric field to the magnitude of the magnetic field must be adjustable to the value required for each particle. In this paper a cyclotron design is considered which would allow operation at a number of discrete frequencies in the 5.5 to 8.6 megacycle range.

Data are presented showing the effect of the physically large variable capacitor upon voltage standing wave, Q, and tuning range. The data have been obtained by making measurements before and after the installation of the parallel plate capacitors. The results indicate that the variable frequency feature is obtained at the expense of Q and voltage gain of the dee stems.

45.2 "Bevatron Operation," by *Dick A. Mack, University of California, Berkeley, Calif.*

Particle accelerators in the billion electron-volt field have substantially extended the range of nuclear research in providing a better understanding of the interactions between high-energy particles. The Bevatron, a proton synchrotron at the University of California Radiation Laboratory, was placed in operation in February 1954. Since that time numerous tests and engineering changes have improved the continuity of operation and the uniformity of accelerated beam pulses. Recently a beam of 3×10^8 protons per pulse has been accelerated to the design energy of 6.3 billion electron-volts (Bev). Refinements in preventive maintenance are in progress to improve the usefulness of the Bevatron as a physics research tool.

45.3 "A 100-Channel Pulse-Height Analyzer Using Magnetic Core Storage," by *Preston W. Byington and C. Wilkin Johnstone, Los Alamos Scientific Lab., Los Alamos, N. M.*

The growing complexity of nuclear experiments often makes the use of a 100-channel pulse-height analyzer highly desirable in order to speed up the taking of data without sacrificing accuracy or resolution. The analyzer to be described uses a linear pulse height to time conversion to achieve equal channel widths. The complete equipment may be subdivided into three groups by function: the pulse-amplitude analyzer, the memory or storage, and the readout.

This paper will be concerned chiefly with storage and readout techniques, but will include a brief discussion of analyzer circuits and performance.

45.4 "One Hundred-Channel Serial Memory Pulse-Height Analyzer," by *T. L. Emmer, Oak Ridge National Lab., Oak Ridge, Tenn.*

When a pulse-height analyzer having in the order of 100 channels is under consideration, the problems of interchannel drift and an economical, trouble-free memory unit for storing pulse-height information are immediately apparent. In the conventional voltage-discriminator type of analyzer, the pulse-height spectrum is divided into channels by a series of biased trigger circuits. Thus any drift in the grid-to-cathode potential of the triggers will cause interchannel drift. Also, each trigger is followed by a series of scalers and a mechanical register to store the pulse-height information. For 100 channels then, this type of storage system is both troublesome and expensive.

In an effort to eliminate these problems, a 100-channel analyzer was designed by G. W. Hutchinson and G. G. Scarrott of Cavendish Laboratories, Cambridge University. Their analyzer employs a mercury delay-line memory unit and a voltage-to-time discriminator.

Information will be given on a second model of this type. This model will incorporate many new features such as a quartz delay line, 120 channels with six times the storage capacity, shorter dead time, better provisions for coincidence counting, brightening of every tenth channel in the display, and provisions for an automatic print-out.

45.5 "Nuclear Reactor Control Systems Utilizing Solid-State Devices," by *Stephen F. Malaker and Edward Rathje, Daystrom Instrument Co., Archbald, Pa.*

The current trend toward industrial participation in the nuclear reactor program with its ultimate goal being economical and competitive production of power has given impetus to the science of control of nuclear power plants. As a consequence, considerable emphasis must be placed upon the design of control instrumentation, since, among other factors, proximity to population centers dictates that, divorced from proper functioning, complete reliability must be an inherent characteristic of any system.

Reliability and rugged long-life features are attributes peculiar to solid-state devices. Recent developments in components and their application in control systems have become widespread. Among the solid-state devices used are thermocouples, rectifiers, transistors, magnetic amplifiers, nonlinear dielectric elements and various combinations of these elements.

The criteria for selecting optimum operating points, dynamic ranges, and specific characteristics will be discussed with emphasis on applications of the diverse solid-state devices to nuclear circuitry, and more specifically to reactor control systems.

45.6 "A New Frequency-Modulation System for the UCRL 184-Inch Cyclotron," by *Quentin A. Kerns, University of California, Berkeley, Calif.*

A modification is planned for the Berkeley 184-inch cyclotron to achieve particle energies commensurate with magnet size. Added magnet coils will raise the flux density to 23 kilogauss and permit the acceleration of protons and deuterons, for example, to energies of 705 and 435 Mev respectively. The higher particle energies however, necessitate greater frequency range (35 to 19 mc for protons, and 17.5 to 13.5 mc for dueterons), while the shorter magnet gap effectively increases the rf energy storage to about one joule. It appeared that these rf specifications precluded the use of the usual rotating capacitor.

SESSION XLVI

THUR. 10:00 A.M.–12:00 Noon

WALDORF-ASTORIA GRAND BALLROOM

Engineering Management II General

Chairman: John F. Byrne, *Motorola, Inc., Riverside, Calif.*

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46.1 "Cost Considerations in Automatic Production," by *E. Finley Carter, Stanford Research Institute, Stanford, Calif.*

New types of automatic production equipment for the electronics manufacturing industry are being developed and installed. Management must decide if the disadvantages of investing capital in new equipment will be offset by such advantages as lower operating cost and higher-value product. The important cost factors which management must consider are discussed. The results of a typical cost analysis are reported. Conclusions drawn from this analysis in regard to production volume, operating costs, and capital investment requirements are discussed.

46.2 "Personal Responsibilities of the Professional Engineer," by *D. J. Simmons, Convair, Ft. Worth, Texas.*

Available engineering manpower can be more effective and each engineer's opportunities for advancement can be increased by recognition and observance of the professional responsibilities of the individual engineer.

Each engineer has responsibilities which he owes to himself, his associates, his supervisor, and his company. If he is a supervisor, he also owes certain responsibilities to those working under his direction. This paper describes these responsibilities.

Each engineer must realize that the primary responsibility for his development and advancement rests entirely upon himself. He must be competent, know how to organize and carry out assignments, and continually develop his abilities.

To his associates, the engineer owes harmonious relations, loyalty, team work, and co-operation.

Observance by the individual engineer of his personal responsibilities will increase his value, effectiveness, and professional stature.

46.3 "The Management of Basic Research," by *T. M. Linville, General Elec. Co., Schnectady, N. Y.*

This paper will describe the objectives and plans of the research that seeks to extend the understanding of materials and forces in nature. It will describe what can be accomplished with people, facilities, money, and other resources.

It will tell how the work and resources can be organized to carry out the objectives and plans; how the resources are to be utilized.

Most importantly, it will tell how the work is done, how it is directed, what stimulates it, what relationships give it strength, and what some of the key human factors are.

It will describe how the work can be evaluated, how it is brought to practical use, and how continual evaluation controls the direction of it.

In addition, it will show how a research laboratory that is focused on finding basic discoveries that widen business horizons and lead to higher living standards can, at the same time, provide staff services to industrial operations and can also provide an extraordinarily valuable "doctor in the house."

46.4 "The Organization and Management of Engineering in a Small Company," by *Roderick M. Scott, Perkin and Elmer, Norwalk, Conn.*

The Perkin-Elmer Corporation is a small corporation by definition for it employs at this moment less than 500 people. It does not intend to remain at its present size, however, for it is a growth organization in every sense. This concept of growth has had a profound influence on the organization of its engineering function.

The group designated as the Engineering Department consists of forty-five trained scientists and engineers; design, drafting and Model Shop personnel to a total of about 115. The Quality Control and Plant Engineering sections, about thirty more, are also a part of our engineering function. Of the many separate problems of management in a group of this description I would like to discuss three in some detail: first, the relative emphasis and control of research and development under direct contract and that for the support and growth of the corporate manufactured product; second, the role of the project engineer and his place in our organization; and third, the mechanism of establishing compensation for engineering personnel.

SESSION XLVII

THUR. 10:00 A.M.–12:30 P.M.

KINGSBRIDGE ARMORY MARCONI HALL

Microwave Theory and Techniques III Ferrites

Chairman: Tore N. Anderson, *Airtron, Inc., Linden, N. J.*

47.1 "Behavior of Ferroxdure at Microwave Frequencies," by *Max T. Weiss, Bell Tel. Labs., Holmdel, N. J.*

Ferroxdure is a ceramic permanent magnetic material with high anisotropy and has a hexagonal crystal structure. Ferromagnetic resonance experiments have been performed on both oriented and unoriented Ferroxdure at various microwave frequencies from 9,000 mc to 72,000 mc. With no externally applied magnetic field a natural resonance occurs near 50,000 mc. With magnetic field applied perpendicular to the hexagonal axis a double resonance occurs for frequencies below 50,000 mc, while with a field applied along the hexagonal axis a single resonance occurs above 50,000 mc. A theory is developed to explain the above results. Significance of these results to the application of Ferroxdure to nonreciprocal microwave circuits will be discussed.

47.2 "Some Applications and Characteristics of Ferrite at Wavelengths of 0.87 CMS and 1.9 CMS," by *Clyde Stewart, Collins Radio Co., Cedar Rapids, Iowa.*

This paper describes the use of ferrites in waveguide to produce Faraday rotations at 0.87 cm and 1.9 cm wavelengths. The Dicke radiometric receiver is briefly reviewed and its improvement by the use of ferrite waveguide components is discussed. Experimental equipment to measure the Faraday rotations consists of a cylindrical section of waveguide, containing ferrite, placed between two transitions from rectangular to circular waveguide. Power from a klystron is applied to the rectangular waveguide and passes through into the circular waveguide. The circular waveguide containing the ferrite is provided with a controllable axial magnetic field which produces a Faraday rotation. The angle of rotation can be measured by rotating the output transition while monitoring the power output. Details are given for the construction of a waveguide unidirectional transmission line for 0.87 cm wavelength.

47.3 "Microwave Devices Using Ferrite and Transverse Magnetic Field," by *Jorgen P. Vinding, Los Gatos, Calif.*

This paper gives a survey of a new class of ferrite devices which have a magnetic field applied transverse to the direction of propagation. Within this class three groups of devices are distinguished: phaseshifters, unidirectional transmission lines, and attenuators.

For each of these the basic principles are briefly discussed and certain design problems are mentioned together with the solutions adopted in particular designs. It is mentioned that other solutions and better results will be possible when better ferrite materials become available.

Among the phaseshifters special attention is given to the single-sideband-modulator and among the unidirectional transmission lines a broadband type is discussed in some detail.

47.4 "Broadband Ferrite Characteristics," by *Murray B. Loss, Sperry Gyroscope Co., Great Neck, N. Y.*

By using two suitably chosen ferrite rods in series in circular waveguide with the magnetic field on one ferrite applied in the opposite direction to that on the other, the frequency sensitivity of Faraday rotation of the combination can be made flat over a large frequency range by the automatic cancellation of the frequency sensitivities of the individual ferrites.

47.5 "Measurement of Microwave Electric and Magnetic Susceptibilities of Ferrite Spheres," by *E. G. Spencer, R. C. LeCraw, and F. Reggia, Diamond Ordnance Fuze Labs., Washington, D. C.*

Resonant cavity methods are described for measuring electric and magnetic susceptibilities of spheres of experimental ferrites of about one millimeter diameter. The Bethe-Schwinger perturbation techniques are used to derive the equations for the susceptibilities in terms of the real and imaginary parts of the frequency shift which are measurable quantities. For a sample placed in the cavity in the position of maximum electric and zero magnetic field, the electric susceptibility is obtained while the converse holds for the magnetic susceptibility. The individual components of the tensor permeability are obtained by measurements using alternately positive and negative circularly polarized waves in the cavity. Data are presented on a magnesium-manganese ferrite to illustrate the techniques described.

SESSION XLVIII

THUR. 10:00 A.M.–12:30 P.M.

KINGSBRIDGE ARMORY FARADAY HALL

Electronic Computers III

Chairman: Ralph E. Meagher, *University of Illinois, Urbana, Ill.*

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48.1 "The Typotron—A Novel Character-Display Storage Tube," by *H. M. Smith, Hughes Aircraft Corp., Culver City, Calif.*

This report describes the Typotron tube, a novel character writing storage tube, developed by the Electron Tube Laboratory of the Hughes Aircraft Company for the Lincoln Laboratory at M.I.T. This tube will display any one or any combination of the 63 characters available on the character matrix incorporated in the writing gun. The characters can be written in less than 50 microseconds, displayed indefinitely until erased at brightness up to a few hundred foot-lamberts. The methods used for character formation, selection, deflection, and display are outlined. Included are schematic diagrams of the Typotron tube and photographs of its essential components and of completed tubes.

48.2 "Electrographic Recording," by *Herman Epstein and Frank T. Innes, Burroughs Corp., Paoli, Pa.*

Some of the properties and possible applications of a new recording technique known as Electrographic Recording include the fact that it is a high-speed technique by which a mark may be put on paper in a duration as small as one microsecond and printed characters, formed from a 5×7 matrix, may be recorded at rates exceeding 5,000 characters per second; for instance, recording rates at hundreds of inches per second of paper strip are potentially possible.

The only motion involved in the technique is the continuously moving paper under the printing head; there are no moving stylii or the like. Further, the system is basically low in cost and consumes very little power; the power necessary to print 5,000 characters per second, serially, requires about 5 watts in addition to that necessary to move the paper.

The system is quiet—essentially the only noise is caused by the moving paper. The printing technique does not involve any messy, wet, or damp processes.

Permanent recording, with no fading, is achieved. No wear of electrodes, or the like, is involved.

In short, this technique offers dry, high-speed, low-cost, low-power, and quiet recording capabilities.

48.3 "Surface-Barrier Transistor Computer Circuits," by *Ralph H. Beter, William E. Bradley, and Ralph B. Brown, Philco Corp., Philadelphia, Pa., and Morris Rubinoff, University of Pennsylvania, Pa.*

The proper use of transistors in computers involves a realization of their special inherent characteristics. Basic computer circuits will be described; that is, flip-flops, multivibrators, and OR and AND gates which result in a minimum number of total components.

48.4 "Semi-Conductor Diode Amplifier Considerations," by *Henry W. Kaufmann, Remington-Rand Inc., Philadelphia, Pa.*

The delay in recovery of back resistance after forward conduction, a phenomenon widely observed in semi-conductor diodes, affords a means of controlling charge at low potential and then using it through a large potential difference. This has been proposed as a means of amplification. Some observations of the relation between forward charge and reverse charge, as functions of time, are reported for samples of commercially available diodes; the significance of these to diode amplification is discussed.

48.5 "An Electronic Circuit for the Generation of Functions of Several Variables," by *Hans F. Meissinger, Reeves Instrument Corp., New York, N. Y.*

Arbitrary functions of two independent variables may be generated electronically by means of diode networks by extending a method commonly used for approximating functions of a single variable. Two variable voltages, x and y, are inserted into the diode network rather than one variable, x, and a fixed bias. This has the effect of making the output of the circuit depend on y as well as on x. The method can be further extended to functions of more than two independent variables.

This article contains a discussion of a systematic design procedure for the function generator and indicates the classes of functions to

which the method is applicable. Illustrative examples of such functions are given. Certain other functions not in this category may be reduced to a suitable form by a preliminary transformation.

The new method eliminates many of the difficulties encountered in representing functions of several variables on electronic analog computers. It offers the advantages of high speed operation, flexibility, and low cost, compared with the methods heretofore in use. The accuracy depends to a large extent on the number of diodes used in the network, and is comparable to that obtained from other function generators employing a bilinear interpolation scheme.

SESSION XLIX

THURS. 2:30–5:00 P.M.

BELMONT-PLAZA MODERNE ROOM

Medical Electronics II General

Chairman: L. H. Montgomery, *Vanderbilt University, Nashville, Tenn.*

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49.1 "New Linear Electron Accelerators for Radiotherapy," by *John C. Nygard and M. G. Kelliher, High Voltage Engrg. Corp., Cambridge, Mass.*

The microwave linear accelerator for the production of intense high-energy electron beams is for the first time now available for general use. Two accelerators with electron energies of 6 and 50 mev will be discussed. The 50-mev electron beam will be utilized directly for cancer therapy, whereas the 6-mev electron beam may be utilized either for the direct sterilization of biological materials or for the production of X-rays for cancer therapy. The general problem of beam and/or patient maneuverability for use in a therapy program, and some aspects of simplified control system requirements will be discussed.

49.2 "Cineradiography," by *Earl R. Miller, Eldon Nickel, and Lee B. Lusted, University of California Medical School, San Francisco, Calif.*

The need for cineradiographic equipment is discussed.

The cineradiographic equipment in the Department of Radiology, University of California School of Medicine, San Francisco, is presented.

The equipment consists of two separate camera units connected to a single X-ray generator.

(a) A 35-mm camera with F0.75 lens and associated control circuits, a Westinghouse Super Dynamax X-ray tube capable of 120KV operation.

(b) A 16-mm camera with F0.85 lens, Philips Image Amplifier and Westinghouse Dynamax 41 X-ray tube.

Each camera unit is connected to the X-ray high voltage by selsyn motors. Film speeds of $1\frac{7}{8}$, $3\frac{3}{4}$, $7\frac{1}{2}$, 15, 30, and 60 frames a second are available.

49.3 "The Use of U-V Microspectrophotographic and Phase and U-V Television Densitometry Techniques in Medical Research," by *Philip O'B. Montgomery, Southwestern Medical School, Dallas, Texas*

The use of microspectrophotographic densitometry techniques in living and nonliving biologic systems with the principal emphasis upon medical research will be illustrated by means of lantern slides of the necessary equipment and the results obtained by the study of arteriolar disease in hypertension and collagen in cancer of the skin.

The use of ultraviolet and phase television, and microspectroscopy and densitometry will be considered from the point of view of research in biologic fields related to normal and abnormal growth. The equipment necessary will be illustrated by means of lantern slides, and preliminary data and projects will be outlined with a consideration of projected general applications in medical research.

49.4 "Application of the Television Ultraviolet Microscope to the Direct Visualization of Cytological Absorption Characteristics," by *George Z. Williams, National Institutes of Health, Bethesda, Md.*

The ultraviolet absorption characteristics of intact living cells are directly visualized on the television screen by the application of industrial television components to the ultraviolet microscope and the use of a specially developed RCA ultraviolet sensitive camera tube. A single line of the cathode-ray picture tube is analyzed by inserting a suitable oscilloscope into the cathode-ray circuit, thus depicting the relative degrees of absorption of the ultraviolet wavelength by the cellular material. Rapid photographic records of selected cells may be simply made.

49.5 "Some Applications of Scanning Techniques in Instrumentation," by *C. Berkley and H. P. Mansberg, Allen B. DuMont Labs., Inc., Clifton, N. J.*

The human may be considered as a data-processing machine supplied with information from transducers of pressure, smell, taste, light, etc., and it is shown that the visual transducer has certain advantages in information handling capabilities. Any information in a visual field can be converted to electronic form by scanning. The resultant waveshape then contains all the information observable visually. Computers can be designed to extract and use this information at a much faster rate than the human. Considerable ingenuity is often required in the information extraction. Some specific applications illustrating typical problems encountered in this process are considered and described, including field scanning and plotting, and optical scanning.

SESSION L

THURS. 2:30–5:00 P.M.

WALDORF-ASTORIA STARLIGHT ROOF

Engineering Management III Symposium

Management Selection as Viewed by Psychologists and Engineering Executives

Chairman: Robert D. Loken, *Life Magazine, New York, N. Y.*

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The selection of management personnel in engineering presents a combination of problems which are unique in the management field. Individually, these problems are approached in several ways.

Psychologists have shown increasing skill in identifying basic traits and abilities. This has led to testing procedures for personnel evaluation.

As various patterns have emerged, extension to managerial areas has followed, and been highly relied upon by some executives. Conversely, countless people in management have reached their positions without psychological evaluation, and they do not rely on these methods for purposes of promotion.

This symposium will explore various approaches to the above problems by psychologists and engineering executives.

50.1 "Selection of Technical Managers as Viewed by a Personnel Psychologist," by *A. P. Johnson, Educational Testing Services, Princeton, N. J.*

The topic of the selection of technical managers can be approached in many ways.

This talk will describe the broad framework and some details of an approach which repeatedly, in other selection situations, has proved effective and which gives promise of being effective in the selection of technical managers.

Basically the approach is simple. For any technical management position find out: (1) the key duties to be performed; (2) the management climate surrounding it; (3) what man under consideration has best demonstrated his ability to do well key duties like these and would be happy in the management climate involved.

50.2 "Psychological Means for the Selection of Managers," by *John C. Flannigan, University of Pittsburgh, Pa.*

50.3 "Selection of Engineering Executives," by *Leron N. Vernon, Personnel Lab., Chicago, Ill.*

Selection of managers for engineering organizations is like selection of other executives with some additional problems. Executive potential in general should be judged in terms of (1) emotional and social maturity, (2) intellectual capacity, and (3) management skills.

Additional problems in selecting engineering executives include the following. (1) An outstanding technologist may earn promotion, but is likely to be a poor manager. The necessary skills, standards, and personal qualities are different. (2) The need for management skills is

greater among engineers because technical men are more difficult to manage than other groups. (3) There is need for a research approach to this problem. Progress will come from better use of available techniques in man evaluation and from scientific study of existing situations. Engineers need to accept and use a science different from their own; in this problem, that science is psychology.

50.4 "Balance in Management Selection," by *Ronald L. McFarlan, Raytheon Mfg. Co., Newton, Mass.*

50.5 "Selection of Technical Management Personnel," by *Dean E. Wooldridge, Ramo-Wooldridge Corp., Los Angeles, Calif.*

The problem of selecting suitable personnel for management of scientific or engineering development work has many elements in common with the problem of selecting personnel for management of other types of activities, but there are also some important differences. One such difference will be particularly emphasized—the importance of selecting successful scientists or engineers, rather than nontechnical people, to manage the activities of other scientists and engineers. The point will be made that, if properly done, the conversion of technically trained men to administrators is by no means the waste of scarce manpower that it is frequently considered to be.

SESSION LI

THURS. 2:30–5:00 P.M.

WALDORF-ASTORIA
ASTOR GALLERY

Electron Devices IV Transistors

Chairman: HARPER Q. NORTH, *Pacific Semiconductors, Inc., Los Angeles, Calif.*

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51.1 "Thermal Properties of Semiconductor Diodes," by *J. N. Carman and W. R. Sittner, Pacific Semiconductors, Inc., Culver City, Calif.*

Experimental and analytical methods for determining steady state and transient thermal properties of semiconductor diodes will be presented. Comparisons of power dissipation constants and thermal time constants of various diode envelope designs will be given.

Curves of measured junction-temperature rise as a function of power dissipation will be shown together with effects of dissipated power on electrical characteristics of typical germanium and silicon diodes.

Operation of silicon and germanium diodes at elevated temperatures and required power derating will be discussed. The relationship between dissipation constants, thermal time constants and high-temperature derating factors will be outlined.

51.2 "Grain Boundaries and Transistor Action," by *Herbert F. Matare, Signal Corps Engrg. Labs., Ft. Monmouth, N. J.*

The advanced knowledge and study of imperfections in crystals has opened a new field of fruitful work. While the "perfect" crystal and its use for electronic devices in the semiconductor field allows *p-n* technique with all its variations as to impurity distribution, the use of defined crystal imperfections, especially in the form of grown grain-boundaries, opens a new variety of possible electronic behavior.

Measurements of grain-boundary orientation in correlation with probe measurements show interesting properties as to current amplification, i.e., release of trapped carriers, and gate-action. The use of grain-boundary-stress fields in transistor electronics should prove also effective in the higher frequency range since these fields which define the characteristics of grain-boundary transistors may be limited to thickness of 1 micron and less.

51.3 "Developments in Silicon Junction Diodes and Power Rectifiers," by *H. Gunther Rudenberg, Transitron Electronic Corp., Melrose, Mass.*

New silicon junction rectifiers are presented. Different alloying techniques provide *p-n* junctions of smaller as well as much larger areas than formerly available. Types developed handle kilowatts of alternating current, others milliwatts of high-frequency signals. They have low forward-voltage drops, excellent rectification efficiencies at rated power, inverse currents of microamperes or fractions thereof, and inverse voltages from 10 to 1,000 volts even at high ambient temperatures. The low-power diodes have minute junctions of 0.1 $\mu\mu$f capacitance, excellent pulse recovery within 0.1 microsecond, cut-off frequencies above 500 mc, and inverse voltages from 10 to 100 volts. High-power rectifiers can operate at currents up to 100 amperes, others at inverse voltages of 1,000 volts. A representative unit measures 5 volts forward drop at 30 amps, and an inverse voltage above 100 volts up to 150 degrees C. Characteristics of types in quantity production will be presented with application data. Thermal design encountered in mounting hermetically-sealed power units will be discussed.

51.4 "Comparative High-Frequency Operation of Junction Transistors Made of Different Semiconductor Materials," by *L. J. Giacoletto, RCA Labs., Princeton, N. J.*

The high-frequency performance of junction transistors is determined in part by the time required for the injected minority carriers to traverse the base region. Large minority carrier mobility contributes to reducing this transit time. However, the hf performance is also affected by factors such as base-lead resistance and collector-to-base junction capacitance, and large majority carrier mobility will reduce the detrimental effects of these factors.

One can determine the design requirements for the optimum hf power amplification. Since both minority and majority carrier mobilities are of about equal importance, it follows that the high-frequency performance of an *n-p-n* transistor should be about the same as that of a geometrically similar *p-n-p* transistor. Furthermore, for evaluation of new semiconductors, a knowledge of the values of both mobilities is required, and a figure of merit is proposed which is formed by the product of the two drift mobilities divided by the square root of the dielectric constant.

51.5 "Characteristics and Some Applications of Fused Junction *P-N-P* Germanium Transistors for High-Frequency Use," by *R. D. Greene, Raytheon Mfg. Co., Newton, Mass.*

The basic characteristics of fused junction *p-n-p* germanium transistors are discussed with particular emphasis on the viewpoint of the circuit designer.

Those parameters of importance in the design of high-frequency transistors are pointed out and optimized. Circuitry for such devices as portable radios, video amplifiers, and gating circuits are discussed.

SESSION LII

THURS. 2:30–5:00 P.M.

WALDORF-ASTORIA
JADE ROOM

Component Parts II General

Chairman: LLOYD T. DEVORE, *General Electric Co., Syracuse, N. Y.*

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52.1 "A Miniature Precision Delay Line," by *James B. Hickey, Rome Air Dev. Center, Griffiss AF Base, Rome, N. Y.*

This paper describes the development, unique construction details, and characteristics of a miniature precision delay line of the *l-c* lumped constant type, having an equivalent "*m*"-derived low-pass filter configuration. This delay line was designed to replace a wire-wound distributed-constant type employed in a complex aircraft identification system. The new lumped constant delay line is superior to the replaced wire-wound distributed constant delay line on the following counts: (1) reproducibility; (2) temperature stability; (3) space factor; (4) availability of delay taps at extremely small delay increments; and (5) versatility of design for a wide range of characteristic impedance and bandwidth.

52.2 "Criteria and Test Procedures for Electromagnetic Delay Lines," by *Norman Gaw and David Silverman, Helipot Corp., Mountainside, N. J.*

The general types of fixed and variable electromagnetic delay lines are reviewed and conventional terminology is defined. Distortions common to delay lines are illustrated and methods for correction are analyzed. Test procedures, both pulse and sinusoidal, and methods for measuring linearity and phase shift are described. The general areas of usefulness of electromagnetic delay lines in today's technologies and some applications are discussed.

52.3 "Evolution of Selenium Rectifier Voltage Ratings," by *N. Bechtold, Signal Corps Engrg. Labs., Ft. Monmouth, N. J.*

The prewar 14-volt plate evolved through accelerated emergency development to the postwar 26-volt plate. Through Signal Corps sponsorship (Contract DA-36-039 SC-107) the 40-volt plate was developed and produced on a large quantity basis. The 1.5 miniaturization factor achieved has been proven reliable in a number of high-voltage power supplies and by virtue of over 15,000 hours of life data. Additional investigations have made use of its advantageous high reverse-resistance for magnetic amplifier applications. Recently, a 60-volt selenium plate was developed under Contract DA-36-039 SC-15454, an extension of the original work. Further advancement of this voltage rating will be contingent upon resolution of the barrier layer stability problem.

52.4 "A Precision Deflection Yoke," by *Harold J. Benzuly, Radio Corp. of Amer., Camden, N. J.*

A method for constructing a deflection yoke assembly suitable for camera pickup tubes using entirely new techniques is described. In particular, a deflection coil assembly for an image orthicon camera is discussed. Although some problems remain to be solved, the electrical and physical characteristics of a standard wound yoke can be realized, making direct substitution with only minor circuit modifications possible.

52.5 "Ceramic-to-Metal Seals for Magnetrons," by *Leo J. Cronin, Raytheon Mfg. Co., Waltham, Mass.*

The advantages of ceramic-to-metal seals are enumerated. Applications of these seals in the construction and operation of magnetrons are discussed and illustrated. The new Mo-Ti sealing technique is described following its historical sequence. Processing techniques such as application of metalizing, firing, and soldering are covered. Mention is made of physical design correlated with stress and strain. Properties of ceramics which were considered when choosing an alumina body are given. The fabrication of ceramics is also mentioned. Results of research utilizing X-ray diffraction and petrographic methods are listed. A selection of slides is presented which shows magnetrons, seals, properties of ceramics and metals, cross sections, equipment, and other salient points.

SESSION LIII

THURS. 2:30–5:00 P.M.

WALDORF-ASTORIA
SERT ROOM

Information Theory III

Chairman: JOHN W. MAUCHLY, *Remington Rand, Inc., Philadelphia, Pa.*

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53.1 "Communication Theory Model and Economics," by *Samuel Bagno, Altertronic Corp., Long Island City, N. Y.*

The mathematical theory of communication is shown to provide a good model for the description of economic activity. Physical production corresponds to transmitting entropy into a communication channel. The economy corresponds to the channel. In the noiseless case, money in circulation corresponds to the capacity of a communication channel. In the presence of noise, maintaining a constant channel capacity requires introducing sufficient redundancy to nullify the equivocation due to noise. The corresponding economic duplication is shown to be the need for an ever-increasing total money supply. The law of diminishing returns and other economic laws have striking analogies in certain theorems of Shannon.

53.2 "Removal of the Redundancy Due to Intersymbol Interference," by *H. Davis, University of California, Los Angeles, Calif., and D. L. Trautman, Hughes Aircraft Co., Culver City, Calif.*

Since the appearance of Shannon's classic paper on communication theory, much attention has been directed to making use of redundancy which might occur in signals in order to reduce the probability of misinterpretations when received in the presence of noise. This has led also to various schemes for using intentionally introduced redundancy to reduce error in reception (error-correcting codes). There are, however, cases where it may be desirable to remove the redundancy present rather than add to it. In this paper we consider discrete parameter, real valued time series which may have members which are statistically dependent (intersymbol redundancy). We transform such series so that the members of the resultant series are mutually independent and such that the resultant series is in one-to-one correspondence with the original signal. This may be useful, for example, if it is desired to transform an arbitrary signal to a Gaussian time series.

53.3 "Noise through Nonlinear Devices," by *Ralph Deutsch, Hughes Aircraft Co., Culver City, Calif.*

A method proposed by Wiener for studying nonlinear transformations of noise voltages is discussed. The technique is extended to a system consisting of a passive linear four-terminal network followed by a member of a general class of nonlinear devices and a linear output filter. The elements are all allowed to interact mutually. For the particular case of bandwidth-limited normal noise, the system equations can be expressed in convenient form. The analysis technique is illustrated by application to a typical physical problem.

53.4 "Linear Filter Optimization with Game Theory Considerations," by *M. C. Yovits and J. L. Jackson, Johns Hopkins University, Silver Spring, Md.*

The optimum reproduction of a signal in noise by a linear filter is considered when the signal spectrum is unknown. The problem is likened to a game; the signal producer by his spectrum choice attempts to *maximize* the filter error while the filter designer attempts to *minimize* it.

The optimum transfer function for any fixed signal spectrum is found; the optimum signal spectrum for any allowable transfer function is found. Their functional intersection gives the game theory solution.

A new variational principle is developed which conveniently gives the *amplitude* of the optimum transfer function for any fixed signal spectrum. Results are listed for some simple cases.

53.5 "The Effect of AGC on Radar Tracking Noise," by *R. H. DeLano and I. Pfeffer, Hughes Aircraft Co., Culver City, Calif.*

The response of the AGC in a radar receiver to the low-frequency fluctuations of the radar echo can produce an increase in the low-frequency spectral density of the angle tracking noise. An increase in noise spectral density by a factor of two to three is representative of what can happen when the envelope of the radar echo is approximately Rayleigh distributed. By simulation, using an electronic analogue computer, the increase in spectral density has been obtained as a function of the product of the envelope bandwidth and the AGC time constant. This consideration must be balanced against other factors requiring a short AGC time constant and one of these is discussed.

SESSION LIV

THURS. 2:30–5:00 P.M.

KINGSBRIDGE ARMORY
MARCONI HALL

Microwave Communications and Systems

Chairman: S. M. KAPLAN, *General Electric Co., Cornell University, Ithaca, N. Y.*

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54.1 "Evaluation of Survey Methods for Use in Microwave Path Analysis," by *W. C. Eddy, Television Associates of Indiana, Inc., Michigan City, Ind.*

There have been many approaches to the problem of evaluating a proposed microwave path. The aerial survey substantiated by a ground survey described in this paper appears best to satisfy the several equations of economy, accuracy and expedition in obtaining data.

This paper describes in detail the techniques that have been developed over the past ten years in aerial survey, as well as outlining in a broad evaluation the other systems that have been used.

The aerial system described in this paper has been used in the original evaluation of many of the presently operating microwave systems. Using data developed by this system, these operating microwave systems have encountered no paths that were marginal or obstructed.

54.2 "A Monopulse Radar Technique," by *R. M. Page, Naval Research Lab., Washington, D. C.*

A technique is described whereby a single radar pulse may provide angle as well as range information for tracking. As a consequence,

angle as well as range information may be presented in the video channel, thus providing simultaneously angle and range information on all echoes independently, so long as they are resolved in range. Since resolution in range is resolution in time, the possibility is suggested that the technique may be applied in radio astronomy to resolve in angle two independent point sources simultaneously in the radio beam, but resolved in time by suitable wide-band circuitry before integration. This is known to work for signals large relative to receiver noise.

54.3 "A Frequency Selective Directional Coupler for Multiplexing," by *Herbert J. Carlin, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

This paper describes the design principles, construction, and operation of a UHF antenna coupler designed to multiplex transmitters and receivers on a single antenna in the frequency band 225–400 mc/sec. The coupler design in first approximation is the realization of a bridged-T circuit in coaxial line employing two tunable precalibrated cavities plus a special balancing network to compensate for antenna mismatch. A more general point of view is to regard the device as a directional coupler with high directivity and with coupling coefficients whose frequency response is similar to that of bandpass and band elimination filters. The resultant multiplexer achieves high isolation and low insertion loss at unusually close frequency spacings between transmitters and receivers. When the multicouplers are used with a frequency separation of 3 mc/sec. (approximately 1 per cent) between euipments, the average insertion loss in a four-coupler system is 1 db per coupler, and the average isolation between equipments located at adjacent couplers is approximately 60 db.

54.4 "Application of Ferrites for Audio Modulation of Microwaves," by *Philip Zirkind, Rome Air Dev. Center, Griffiss AF Base, Rome, N. Y.*

A novel method of transmission of audio frequencies by means of microwaves is described. The plan of polarization of a microwave carrier of 9,375 mc is rotated in varying amounts when propagated through a ferrite inserted in a circular waveguide and subjected to a longitudinal magnetic field varying in strength at audio frequencies. Rectangular waveguides which are perpendicular to each other are coupled to both sides of the ferrite section. It is then observed that in the waveguide beyond the ferrite section in the direction of propagation, the microwave carrier is modulated in proportion to the variations in the magnetic field strength applied to the ferrites. Further observations are described of the propagation of this modulated microwave carrier through space, and detection at a receiver. The means of demodulation and the characteristics of the demodulated signal are described.

54.5 "A Narrow-Band Radar Relay System," by *C. W. Doerr and J. L. McLucas, Haller, Raymond and Brown, Inc., State College, Pa.*

The HRB Narrow-Band Radar Relay System provides for presenting at a remote point the PPI display from any standard surveillance or airport radar which uses a continuously rotating antenna.

Basic units are the encoder, the decoder, and the indicator. The encoder first reduces the bandwidth of the radar video to the audio range through a process of integration. The encoder also generates a rate signal, whose frequency is related to antenna rotation speed, and a North mark. These three signals are superimposed on subcarriers and mixed before being fed to a standard communications link. The decoder separates the three signals and uses them to create PPI display on Indicator.

SESSION LV

THURS. 2:30–5:00 P.M.

KINGSBRIDGE ARMORY FARADAY HALL

Aeronautical and Navigational Electronics III Navigation

Chairman: GEORGE RAPPAPORT, *Aircraft Radiation Lab., Wright Field, Dayton, Ohio*

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55.1 "An All-Weather Radio Sextant," by *D. O. McCoy, Collins Radio Co., Cedar Rapids, Iowa*

The techniques used in radio astronomy have been applied to produce an all-weather solar tracking instrument for navigational purposes. The principle of operation of this device is described. An illustrated description of a typical radio sextant is given together with a discussion of the accuracies obtained from several versions of the equipment.

55.2 "A UHF Ground-Based Automatic Direction Finder," by *Robert L. Cattoi, Collins Radio Co., Cedar Rapids, Iowa*

This paper presents a general description of the Collins DF-101 ground-based automatic direction finder. The described system gives a continuous bearing indication of the relative direction of arrival of amplitude modulated or unmodulated signals in the 225–400 megacycle frequency range. Design goals included over-all simplicity, relatively small size, portability and good operational accuracy. The paper outlines design requirements, brief operational theory, development problems, physical and electrical description, as well as equipment application and performance.

55.3 "Wullenweber-Type Ultra High-Frequency Radio Direction Finder," by *Richard C. Benoit, Jr., Rome Air Dev. Center, Griffiss AF Base, Rome, N. Y., and W. M. Furlow, Jr., Melpar, Inc., Alexandria, Va.*

This paper deals with the development and fabrication of an ultra high-frequency Wullenweber type of radio direction finder capable of a 10 to 1 frequency coverage at a high order of azimuth bearing resolution permitting the display of simultaneous bearings on a single frequency.

It is primarily designed to cover the 225–400 mc band but also offers coverage from 100–1,000 mc at a possible sacrifice of performance. The antenna system consists of 100 vertically polarized dipoles having an array diameter of 10 meters which are utilized to produce a beamwidth in the order of 6 degrees sweeping in azimuth at 600 rpm.

Of the several components comprising the system, detailed discussion is given of the radio frequency switch, couplers, and phasing elements. The switch was designed to connect simultaneously each of 36 adjacent dipoles to 36 phasing lines, using transformers made of flat strip transmission lines to eliminate mechanical contacts. Cross-talk between channels and insertion loss are negligible. The coupler matches 36 inputs of 50 ohms each to a 50-ohm output cable and is made up of five cascaded coaxial transformers, telescoped together and loaded with dielectric material to minimize over-all size.

Aperture Wullenweber systems are also presented.

55.4 "High-Precision Computer for Automatic Solution of the Celestial Triangle," by *Gene R. Marner, Collins Radio Co., Cedar Rapids, Iowa*

A precise electromechanical analog computer which automatically solves the celestial triangle will be described. The inputs are shaft positions representing the latitude of the instrument and the declination and local hour angle of a point on the celestial sphere. The computer then continuously presents shaft positions representing the altitude and azimuth of the point in question. No restrictions are placed on the range of any of the angles involved. The results of an extensive experimental error analysis will be presented, and applications to navigation systems and to the direction of altitude-azimuth mounted radio telescopes will be pointed out.

55.5 "JAINCO: a High-Precision Lightweight Aircraft Navigational System," by *Donald H. Jacobs, Jacobs Instrument Co., Bethesda, Md.*

JAINCO is a high-precision phase comparison type of navigational system involving three ground stations: a master and two slaves. Each ground station transmits five sinusoidal modulation frequencies on its carrier. Phase comparisons are made in the airplane automatically and simultaneously on all these modulations. These comparisons give two unambiguous hyperbolic coordinates of position with six-figure accuracy. One set of ground stations can control an unlimited number of aircraft. Other important characteristics are the small size and lightness of ground and airborne equipments, the fact that no airborne operator is required, and the fact that the system gives unambiguous position readings (no cycle counting). The aircraft, being guided, emits no signals.

IRE News and Radio Notes

A. N. Goldsmith Made Eminent Member of Eta Kappa Nu Honor Society

The fifty-six college chapters of Eta Kappa Nu, the electrical engineering honor society, have elected Alfred N. Goldsmith to the rank of Eminent Member. Induction ceremonies were held on January 31 at the Henry Hudson Hotel. Following the ceremonies, new Eminent Members were presented at Eta Kappa Nu's Award Dinner.

Over the years only twenty men have achieved Eminent Membership, which is offered "to those individuals who by their technical attainments and contributions to society have shown themselves to be outstanding leaders in the field of electrical engineering and great benefactors of their fellow men."

Buenos Aires Section Holds Eighth Convention

The Buenos Aires Section held its eighth annual convention in November. The program presented technical papers, tours, an exhibition of communications and electronic equipment, and a banquet.

Technical sessions included the following papers: G. Von Trentini, "Radar Antennae"; P. J. Noizeux, "The History of Communications in Argentina"; S. Castro, "The Future of Electronics in Our Laboratories"; J. C. Coriat, "Manufacture of Television Receivers in Argentina."

An exhibition held in the Argentine Automobile Club featured the latest electronic equipment designed and manufactured in Argentina. P. N. Guzzi, Chairman of the Buenos Aires Section, presided at inauguration ceremonies. In his speech to the group, Mr. Guzzi described the part taken by the Buenos Aires Section in the development of electronics in Argentina.

Tours were conducted at the Transradio International, largest transmission station in the country, at Compañia Standard Electric Argentina, and at Capehart Argentina.

ASEE Will Meet in June

The annual meeting of the American Society for Engineering Education will be held June 20–24, 1955, at Pennsylvania State University, State College, Pennsylvania. The meeting will be attended by over 2,000 educators and educational administrators from colleges throughout the country. The society, this year especially welcoming foreign visitors, will present a program featuring international aspects of engineering education.

The program will offer the reports of two committees. The Committee on Evaluation of Engineering Education will report on the curricular goals of engineering education, emphasizing changes needed to keep engineering education abreast of scientific and technological advances. A second committee will discuss the teaching of humanities, social sciences, and business studies in the engineering curriculum.

The Engineering College Research Council will have a session dealing with education for research, while the Engineering College Administrative Council will feature a session on methods of interesting secondary school students in engineering as a career. Over 75 other conferences, concerning all areas of engineering education as well as related areas, will be offered.

The printed program for the meeting will be available in May, and further information may be obtained from the Secretary of the ASEE, Northwestern University, Evanston, Illinois.

New Appointments Made to the Board of Directors

At the meeting of the IRE Board of Directors on January 6 in New York, a number of appointments to the Board of Directors were made for this year. W. R. G. Baker was appointed Treasurer; Haraden Pratt, Secretary; J. R. Pierce, Editor; A. N. Goldsmith, Director; A. V. Loughren, Director; and Howard Vollum, Director.

P. N. Guzzi, Chairman of the Buenos Aires Section, speaks to the Convention group on recent developments made in the field of electronics in Argentina and on the contribution of the Buenos Aires Section to this area.

Calendar of Coming Events

IRE Philadelphia Section Color TV Symposium, Physics Building, University of Pennsylvania, March 8, 15, 22, 29, and April 5

Symposium on Electromagnetic Relays, Oklahoma A and M College, Stillwater, Oklahoma, March 9–11

IRE National Convention, Waldorf-Astoria Hotel and Kingsbridge Armory, New York, N. Y., March 21–24

IRE-PIB Symposium on Modern Network Synthesis, Engineering Societies Building, New York, N. Y., April 13–15

Instrumentation Symposium, Proving Ground Instrumentation Committee of the American Ordnance Association, Patrick Air Force Base, Cocoa, Florida, April 14 and 15

SMPTE 77th Semiannual Convention, Hotel Drake, Chicago, Ill., April 17–22

International Symposium on Electrical Discharges in Gases, Technical University, Delft, Netherlands, April 25–30

IRE Seventh Region Technical Conference and Trade Show, Hotel Westward Ho, Phoenix, Ariz., April 27–29

Semiconductor Symposium, Electrochemical Society, Cincinnati, Ohio May 2–5

National Aeronautical Electronics Conference, Biltmore Hotel, Dayton, Ohio, May 9–11

IRE-AIEE-IAS-ISA National Telemetering Conference, Hotel Morrison, Chicago, Ill., May 18–20

IRE-AIEE-RETMA-WCEMA Electronic Components Conference, Hotel Ambassador, Los Angeles, Calif., May 26–27

American Society for Engineering Education Annual Meeting, Pennsylvania State University, State College, Pennsylvania, June 20–24

URSI-U. of Michigan International Symposium on Electromagnetic Wave Theory, University of Michigan, Ann Arbor, Mich., June 20–25

International Analogy Computation Meeting, Societe Belge des Ingenieurs des Telecommunications et D'Electronique, Brussels, Belgium, September 27–October 1

IRE-AIEE Conference on Industrial Electronics, Rackham Memorial Building, Detroit, Michigan, September 28–29

Second International Automation Exposition, Chicago Navy Pier, Chicago, Illinois, November 14–17

Convention Social Program to Feature Gen. Ridgway

The National Convention activities (p. 347) will include an outstanding program of social events and women's activities.

Social Events

A get-together cocktail party will be held from 5:30 P.M. to 7:30 P.M. on March 21 in the Grand Ballroom of the Waldorf. Tickets may be purchased at the door.

The Annual IRE Banquet on March 23 will also be held in the Waldorf's Grand Ballroom. The featured speaker will be General Matthew B. Ridgway, Chief of Staff, U. S. Army. The winners of the awards for 1955 will be honored during the banquet in ceremonies conducted by President John D. Ryder. Reservations (while they last) may be purchased in advance from IRE at $14.00 apiece.

Women's Program

Registration and headquarters for women's activities will be held in the Regency Suite on the fourth floor of the Waldorf. In addition to the program outlined below, tickets to radio and TV broadcasts will be available gratis.

Monday, March 21

9:30 A.M.–5:00 P.M.—Registration—Waldorf-Astoria, Regency Suite, Fourth Floor. Get-Together Coffee Hour, 9:30 A.M.–10:30 A.M.

1:45 P.M.–4:30 P.M.—Behind the Scenes at R.H. Macy Department Store (a view of merchandising from manufacturer to customer).

Tuesday, March 22

9:15 A.M.–11:00 A.M.—Conducted tour of European Masterpieces, Metropolitan Museum of Art. (One-way transportation included.) @ $1.75

11:30 A.M.–1:00 P.M.—Luncheon (Optional) Metropolitan Museum Cafeteria Restaurant. (Price of luncheon extra, as ordered)

3:00 P.M.–5:00 P.M.—Tea at new IRE Headquarters Building, 5 East 79 Street.

Wednesday, March 23

8:15 A.M.–11:30 A.M.—Tour of United Nations, including a meeting of the "Commission on the Status of Women." @ $1.50

12:00 NOON–2:00 P.M.—Luncheon and Fashion Show, Empire Room, Waldorf-Astoria Hotel. @ $4.50.

2:30 P.M. Matinee. "Kismet" or "Teahouse of the August Moon." @$4.55

Thursday, March 24

8:50 A.M.–11:00 A.M.—N.B.C. Tour of Radio City, 30 Rockefeller Plaza, including "Break the Bank" radio broadcast. @ $.70

URSI Invites Papers for Spring Meeting

The URSI Spring Meeting will be held at the National Bureau of Standards in Washington, May 3 to 5. The meeting will be sponsored by the USA National Committee of the International Scientific Radio Union and co-sponsored by the PG on Antennas and Propagation and the PG on Microwave Theory and Techniques. The meeting will include a combined technical session followed by separate sessions in various fields.

Authors who wish to present papers are asked to submit titles and one or two hundred word abstracts by March 1. Papers should be submitted to the representative in each field as follows: Radio Measurements and Standards, *Dr. R. G. Fellers, Bldg. 3, PH, Radio 1, Naval Research Laboratory, Washington 20, D. C.;* Radio and Troposphere, *Dr. J. B. Smyth, U. S. Navy Electronics Laboratory, San Diego 52, Calif.;* Ionospheric Radio, *Dr. Millett Morgan, Thayer School of Engineering, Dartmouth College, Hanover, New Hampshire;* Radio Noise of Terrestrial Origin, *F. H. Dickson, Communications Liaison Branch, Room 2D276, National Defense Building, Washington 25, D. C.;* Radio Astronomy, *Dr. J. P. Hagan, Naval Research Laboratory, Washington 20, D. C.;* Radio Wave and Circuits, *Dr. E. C. Jordan, Dept. of E.E., U. of Ill., Urbana, Ill.;* Radio Electronics, *Dr. W. G. Shepherd, Inst. of Tech., U. of Minn., Minneapolis 14, Minnesota;* PG on Antennas and Propagation, *H. F. Englemann, Federal Telecommunication Labs., Nutly, N. J.;* Microwave Theory and Technique, *J. T. Booljahn, Stanford Research Institute, Stanford, California.*

Proceedings of Electronic Components Symposium Now Available

The *1954 Electronic Components Symposium Proceedings* are available at $4.50 per copy. Those interested in obtaining *The Proceedings* may write to A. E. Zdobysz, Associated Engineering Services, Incorporated, 1 Thomas Circle, N.W., Washington 5, D. C.

Ralph Hackbusch (*right*), former IRE Vice-President, receiving radio citation from G. C. W. Browne.

Ralph Hackbusch Receives Citation Upon Retirement

Ralph A. Hackbusch, Vice-President of the IRE in 1944, has recently retired as President of the Canadian Radio Technical Planning Board. G. C. W. Browne, Controller of Telecommunications of the Department of Transport, presented Mr. Hackbusch, upon his retirement, with a citation. The award was made in "recognition of his enterprise and leadership in formulating sound engineering principles and organizing technical facts to assist in the development of the Canadian radio industry and radio service of the Nation."

In addition to his position on the planning board, Mr. Hackbusch has been a member of the Canadian Electrical Code Committee, the Canadian Standards Association, and the Radio Manufacturers Association.

Brig. Gen. B. S. Kelsey to Make Chief Speech at PGANE Luncheon in New York

Brigadier General Benjamin S. Kelsey, USAF, will be the principal speaker at the annual luncheon of the Professional Group on Aeronautical and Navigational Electronics to be held March 22 during the National Convention in New York. General Kelsey, who formerly headed the USAF's All-Weather Flying division, will speak on "The First and Last Thousand Feet." Gen. Kelsey's present post is Deputy Director of Research and Development, Office of the Deputy Chief of Staff for Development.

The PGANE luncheon will be held in the Empire Room of the Waldorf-Astoria, site of two PGANE-sponsored technical sesisons to be held the same day. Tickets, priced at $3.50, may be obtained by writing to William P. McNally, W. L. Maxson Corporation, 460 West 34 Street, New York 1, New York.

Telemetering Group to Meet

The National Telemetering Conference, under the sponsorship of the IRE (Radio Telemetering and Remote Control Professional Groups participating), American Institute of Electrical Engineers, Instrument Society of America, and the Institute of Aeronautical Sciences, will be held at the Morrison Hotel in Chicago, May 18–20. Kipling Adams of the General Radio Company is National Chairman. Thirty-six papers will be presented at the conference, and the following schedule of papers has been planned:

Wednesday, May 18

Morning

Flight Control

Afternoon

System I—Industrial Telemetering
Components I—Pickups and Transducers

Thursday, May 19

Morning

Systems II—Flight Testing

Afternoon

Systems III—Data Processing
Components II—Transmitting Link

Friday, May 20

Morning

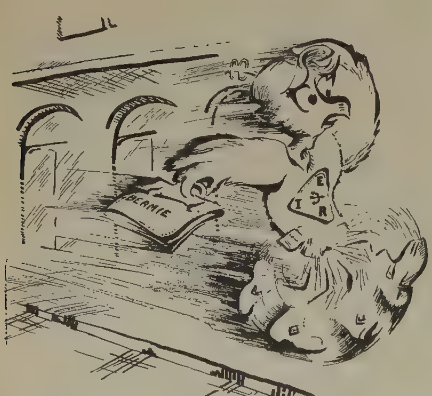
Systems IV—New Developments in Telemetering and Remote Control

Afternoon

Components III—Data Processing

Executive Committee Approves Two New Chapters

At its meeting of January fifth, the Executive Committee approved two P. G. Chapters. They are: Microwave Theory and Techniques, Northern New Jersey Chapter; Production Techniques, Washington Chapter. The Pasadena Subsection of the Los Angeles Section was also approved.



This is Beamie, a character who will appear frequently in the new IRE Student Quarterly, which goes to student members without extra charge, and to which other IRE members can subscribe for $2.00 a year. Here Beamie is shown traveling with the speed of light at the IRE Convention, trying to see it all on his lunch hour. (But he didn't make it.)

TRANSACTIONS OF IRE PROFESSIONAL GROUPS

The following issues of TRANSACTIONS are available from the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y., at the prices listed below.

Sponsoring Group	Publications	Group Members	IRE Members	Non Members*
Aeronautical & Navigational Electronics	PGAE-5: A dynamic Aircraft Simulator for Study of Human Response Characteristics (6 pages)	$ .30	$ .45	$ .90
	PGAE-6: Ground-to-Air Co-channel Interference at 2900 MC (10 pages)	.30	.45	.90
	PGAE-8: June 1953 (23 pages)	.65	.95	1.95
	PGAE-9: September 1953 (27 pages)	.70	1.05	2.10
	Vol. ANE-1, No. 1: March 1954 (51 pages)	1.00	1.50	3.00
	Vol. ANE-1, No. 2: June 1954 (22 pages)	.95	1.40	2.85
Antennas and Propagation	PGAP-4: IRE Western Convention, August 1952 (136 pages)	2.20	3.30	6.60
	Vol. AP-1, No. 1: July 1953 (30 pages)	1.20	1.80	3.60
	Vol. AP-1, No. 2: October 1953 (31 pages)	1.20	1.80	3.60
	Vol. AP-2, No. 1: January 1954 (39 pages)	1.35	2.00	4.05
	Vol. AP-2, No. 2: April 1954 (41 pages)	2.00	3.00	6.00
	Vol. AP-2, No. 3: July 1954 (36 pages)	1.50	2.25	4.50
	Vol. AP-3, No. 4: October 1954 (36 pages)	1.50	2.25	4.50
Audio	PGA-5: Design Interrelations of Records and Reproducers (8 pages)	.30	.45	.90
	PGA-7: Editorials, Technical Papers, and News, May 1952 (47 pages)	.90	1.35	2.70
	PGA-10: November–December 1952 (27 pages)	.70	1.05	2.10
	Vol. AU-1, No. 1: January–February 1953 (24 pages)	.60	.90	1.80
	Vol. AU-1, No. 2: March–April 1953 (34 pages)	.80	1.20	2.40
	Vol. AU-1, No. 3: May–June 1953 (34 pages)	.80	1.20	2.40
	Vol. AU-1, No. 4: July–August (19 pages)	.70	1.05	2.10
	Vol. AU-1, No. 5: September–October 1953 (11 pages)	.50	.75	1.50
	Vol. AU-1, No. 6: November–December 1953 (27 pages)	.90	1.35	2.70
	Vol. AU-2, No. 1: January–February 1954 (38 pages)	1.20	1.80	3.60
	Vol. AU-2, No. 2: March–April 1954 (31 pages)	.95	1.40	2.85
	Vol. AU-2, No. 3: May–June 1954 (27 pages)	.95	1.40	2.85
	Vol. AU-2, No. 4: July–August 1954 (27 pages)	.95	1.40	2.85
	Vol. AU-2, No. 5: September–October 1954 (22 pages)	.95	1.40	2.85
Broadcast and Television Receivers	PGBTR-1: Round-Table Discussion on UHF TV Receiver Considerations, 1952 IRE National Convention (12 pages)	.50	.75	1.50
	PGBTR-3: June 1953 (67 pages)	1.40	2.10	4.20
	PGBTR-5: January 1954 (96 pages)	1.80	2.70	5.40
	PGBTR-6: April 1954 (119 pages)	2.35	3.50	7.00
	PGBTR-7: July 1954 (58 pages)	1.15	1.70	3.45
	PGBTR-8: October 1954 (20 pages)	.90	1.35	2.70
Circuit Theory	PGCT-1: IRE Western Convention, August 1952 (100 pages)	1.60	2.40	4.80
	PGCT-2: Papers presented at the Circuit Theory Sessions of the Western Electronic Show and Convention, San Francisco, California, August, 19–21, 1953, (106 pages)	1.95	2.90	5.85
	Vol. CT-1, No. 1: March 1954 (80 pages)	1.30	1.95	3.90
	Vol. CT-1, No. 2: June 1954 (39 pages)	1.00	1.50	3.00
	Vol. CT-1, No. 3: September 1954 (73 pages)	1.00	1.50	3.00
	Vol. CT-1, No. 4: December 1954 (42 pages)	1.00	1.50	3.00
Communications Systems	Vol. CS-2, No. 1: January 1954 (83 pages)	1.65	2.50	4.95
Component Parts	PGCP-1: March 1954 (46 pages)	1.20	1.80	3.60
	PGCP-2: September 1954: Papers presented at the Component Parts Sessions at the 1954 Western Electronic Show and Convention, Los Angeles, California (118 pages)	2.25	3.35	6.75
Electronic Computers	Vol. EC-2, No. 2: June 1953 (27 pages)	.90	1.35	2.70
	Vol. EC-2, No. 3: September 1953 (27 pages)	.75	1.10	2.25
	Vol. EC-2, No. 4: December 1953 (47 pages)	1.25	1.85	3.75
	Vol. EC-3, No. 1: March 1954 (39 pages)	1.10	1.65	3.30
	Vol. EC-3, No. 2: June 1954 (65 pages)	1.65	2.45	4.95
	Vol. EC-3, No. 3: September 1954 (54 pages)	1.80	2.70	5.40
	Vol. EC-3, No. 4: December 1954 (46 pages)	1.10	1.65	3.30

* Public libraries, colleges, and subscription agencies may purchase at IRE member rate.

(*Continued on the following page*)

TRANSACTIONS OF IRE PROFESSIONAL GROUPS

(Continued)

Sponsoring Group	Publications	Group Members	IRE Members	Non-Members*
Electron Devices	PGED-2: Papers on Electron Devices presented at the IRE National Conference on Electron Tube Research, Ottawa, Canada, June 16–17, 1952, and IRE Western Convention, Long Beach (84 pages)	$1.60	$2.40	$4.80
	PGED-4: December 1953 (62 pages)	1.30	1.95	3.90
	Vol. ED-1, No. 1: February 1954 (72 pages)	1.40	2.10	4.20
	Vol. ED-1, No. 2: April 1954 (75 pages)	1.40	2.10	4.20
	Vol. ED-1, No. 3: August 1954 (77 pages)	1.40	2.10	4.20
Engineering Management	PGEM-1: February 1954 (55 pages)	1.15	1.70	3.45
	PGEM-2: November 1954 (67 pages)	1.30	1.95	3.90
Industrial Electronics	PGIE-1: August 1953 (40 pages)	1.00	1.50	3.00
Information Theory	PGIT-2: A Bibliography of Information Theory (Communication Theory-Cybernetics) (60 pages)	1.25	1.85	3.75
	PGIT-3: March 1954 (159 pages)	2.60	3.90	7.80
	PGIT-4: September 1954 (234 pages)	3.35	5.00	10.00
Instrumentation	PGI-2: Data handling Systems Symposium: IRE Western Electronic Show & Convention, Long Beach, California, August 27–29, 1952 (111 pages)	1.65	2.45	4.95
	PGI-3: April 1954 (55 pages)	1.05	1.55	3.15
Medical Electronics	PGME-1: November 1953 (40 pages)	1.05	1.55	3.15
Microwave Theory and Techniques	Vol. MTT-1, No. 2: November 1953 (44 pages)	.90	1.35	2.70
	Vol. MTT-2, No. 1: Papers presented at the Microwave Radio Relay Systems Symposium, New York N. Y., April 1954 (107 pages)	2.20	3.30	6.60
	Vol. MTT-2, No. 2: July 1954 (67 pages)	1.25	1.85	3.75
	Vol. MTT-2, No. 3: September 1954: Papers presented at the joint IRE Professional Group-URSI meeting held in Washington, D. C., May 5, 1954 (54 pages)	1.10	1.65	3.30
Nuclear Science	Vol. NS-1, No. 1: September 1954 (42 pages)	.70	1.00	2.00
Reliability and Quality Control	PGQC-1: Papers presented at the 1951 Radio Fall Meeting, and 1952 IRE National Convention (58 pages)	1.20	1.80	3.60
	PGQC-2: March 1953 (51 pages)	1.30	1.95	3.90
	PGQC-3: February 1954 (39 pages)	1.15	1.70	3.45
Telemetry and Remote Control	PGRTRC-1: August 1954 (16 pages)	.85	1.25	2.55
Ultrasonics Engineering	PGUE-1: June 1954 (62 pages)	1.55	2.30	4.65
	PGUE-2: November 1954 (43 pages)	1.05	1.55	3.15
Vehicular Communications	PGVC-2: Symposium on What's New in Mobile Radio (32 pages)	1.20	1.80	3.60
	PGVC-3: Theme: Spectrum Conservation Washington, D. C., December 3–5, 1952 (140 pages)	3.00	4.50	9.00
	PGVC-4: Design, Planning and Operation of Mobile Communications Systems, June, 1954 (98 pages)	2.40	3.60	7.20

* Public libraries, colleges, and subscription agencies may purchase at IRE member rate.

Technical Committee Notes

The **Facsimile** Committee met at the Times Annex on December 10 with Vice Chairman A. G. Cooley presiding. The test chart and definitions being prepared by the committee were discussed.

The **Audio Techniques** Committee met at IRE Headquarters on December 15 with Chairman D. E. Maxwell presiding. The work of three new subcommittees was discussed and further revisions were made in the proposed *Measurements Standard* (on amplification, bridging loss, and insertion loss; power loss and transducer loss; attenuation; transformer loss and transition loss).

The **Standards** Committee met at IRE Headquarters on January 13 with Vice-Chairman Baldwin presiding in the morning and Chairman Weber presiding during the afternoon session. The committee discussed procedure for IRE delegates in voting on ASA Standards, and considered two groups of definitions of ASA Sectional Committee C42. The committee voted to request that the IRE Executive Committee set up a Technical Committee on Radio Frequency Interference. The proposed *Standards on Antennas and Waveguides: Definitions for Waveguide Components* were reviewed and conflicts with ASA C42 definitions were discussed at length.

The **Ad Hoc Committee on Terminology in Spurious Radiation Work** met January 12 with Chairman Wayne Mason presiding. Problems in terminology were discussed and definition for "extraneous emission" was tentatively selected.

The **Ad Hoc Committee on Spurious Radiation** met on January 12 with R. F. Shea, Chairman, presiding. Mr. Shea described the tri-service sponsored conference on interference held in December in Chicago by the Armour Research Foundation. Mr. Glenn told of the work being done on open field measurements and screen room measurements by IRE Subcommittee 17.4. Mr. Cumming announced that an Interference Symposium has been planned for the 1955 IRE Convention.

The committee voted to request the formation of a new technical committee on radio frequency interference. Work on interference done by Roy Abbett was not assigned to a technical committee at this time, so that developments on the new committee could be considered.

The **Radio Transmitters** Committee met at IRE Headquarters on January 12 with Chairman P. J. Herbst presiding. The committee discussed definitions for terms in the field of spurious output. The committee considered the following proposed standards now being prepared by the various subcommittees: proposed standards on pulses; methods of measurement of pulse quantities; proposed standards on monochrome television transmitters; and proposed methods of testing radio-telegraph transmitters (below 50 mc). The committee also discussed IRE published standards on radio transmitters which were in need of review and revision.

The **Navigation Aids** Committee met on November 19 and on December 8 at IRE Headquarters, and H. R. Mimno presided. The committee continued work on the proposed *Standards on Direction Finder Measurements*.

The **Video Techniques** Committee met at IRE Headquarters on November 29 with Chairman W. J. Poch presiding. The committee discussed the IRE-RETMA spurious radiation program. Vice-chairman J. L. Jones reported on the recent progress made on the measurement of "differential gain" and "differential phase."

P. H. Smith presided when the **Antennas and Waveguides** Committee met on November 17 at IRE Headquarters. The Committee discussed the proposed *Definitions of Waveguide Components* and the proposed *Navigation Aids Definitions*. It was noted that the Measurements of *Waveguide Components* (in preparation) should be restricted to linear and reciprocal elements, and that the advice of the Circuits Committee should be requested on the use of "port" nomenclature.

On December 1, the **Industrial Electronics** Committee met at IRE Headquarters, with Chairman J. E. Eiselein presiding. The committee decided that the Professional Group on Production Techniques should be notified of their desire to originate and sponsor IRE standard definitions, standard methods of test and measurements, and good engineering practices relating to automation, including automatic machinery for production of electronic equipment.

Abstracts of Transactions of the I.R.E.

The following issues of "Transactions" have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
Antenna and Propagation	Vol. AP-3, No. 1	$1.60	$2.40	$4.80
Broadcast and Television Receivers	Vol. BTR-1, No. 1	$1.25	$1.85	$3.75
Electron Devices	Vol. ED-1, No. 4	$3.20	$4.80	$9.60

* Public libraries and colleges may purchase copies at IRE Member rates.

ANTENNAS AND PROPAGATION

Vol. AP-3, No. 1, January, 1955

News and Views

Double Parabolic Cylinder Pencil-Beam Antenna—R. C. Spencer, F. S. Holt, H. M. Johanson, and J. Sampson

Radiation from a point source placed on the focal line of a parabolic cylinder is reflected in succession from this cylinder and from a second parabolic cylinder crossed so that its focal line coincides with the directrix of the first cylinder. The two reflections result in a parallel beam. The theory is applicable to both microwaves and light. The advantages of shipping the cylinders in the form of flat sheets and the possibilities of independent control of horizontal and vertical beamwidths and shapes are pointed out. Experimental models have been built and tested.

An Atmospheric Analyzer—P. F. Smith

The enormously powerful radiation from lightning discharges has been used as a tool for the study of the ionosphere at very low frequencies. Previous workers have calculated the height of the ionosphere and the distance to the radiation source by application of the ray theory of multiple ionospheric reflections. Time measurements for use in the height and distance formulas were obtained by direct measurement of recorded oscillograms. Time resolution was limited and the calculations were extremely tedious.

This paper describes a complete system which records the received atmospheric wave form and simultaneously derives a plot of the time intervals between peaks of the wave form vs time. The time-interval plot greatly facilitates data reduction by means of a simple curve-fitting technique. Accuracy of measurement of height and distance is limited chiefly by the validity of the multiple reflection theory and the assumption of a simple layer structure in the ionosphere. The time differences are measured more accurately than is ordinarily possible with direct measurement on the waveform. The system automatically derives frequency-time plots for whistling atmospherics and relates the whistler to preceding impulsive atmospherics. Consumption of photographic film is low. The system can be used to visually monitor tape-recorded atmospherics.

A Single-Control Tuning Circuit for Electrically Small Antennas—R. E. Webster

It is often desirable to tune electrically small antennas at their feed terminals. An "L" matching circuit for this purpose is considered. It is shown that one fixed element and one variable tuning element are adequate for certain small antennas. The manner in which losses affect the circuit efficiency and the single-control tuning feature is discussed. Application of the technique to a tunable cavity antenna is noted.

Design of Line-Source Antennas for Narrow Beamwidth and Low Side Lobes—T. T. Taylor

It is well known that the phenomenon of radiation from line-source antennas is very similar to that of the diffraction of light from narrow apertures. Unlike the optical situation, however, antenna design technique permits the use of other-than-uniform distributions of field across the antenna aperture. Line source synthesis is the science of choosing this distribution function to give a radiation pattern with prescribed properties such as, for example, narrow angular width of the main lobe and low side lobes. In the present article the mathematical relationships involved in the radiation calculation are studied from the point of view of function theory. Some conclusions are drawn which outline the major aspects of synthesis technique very clearly. In particular, the problem of constructing a line source with an optimum compromise between beamwidth and side-lobe level (analogous to the Dolph-Tchebycheff problem in linear array theory) is considered. The ideal pattern is cos $\pi\sqrt{A^2-u^2}$, where $u=(2a/\lambda)$ cos θ, a is the half-length of the source, and cosh πA is the side-lobe ratio. Because of theoretical limitations, this pattern cannot be obtained from a physically realizable antenna; nevertheless its ideal characteristics can be approached arbitrarily closely. The procedure for doing this is given in detail.

On the Input Conductance of Thin Antennas—G. Barzilai

Input conductances of thin antennas are computed by evaluating separately input voltage and radiated power for a prescribed value of the current near the maximum. The current distribution required to evaluate the radiated power and the input voltage is obtained by solving Hallèn's equation for the prescribed value of the current. First and second order computations were carried out for full wave antennas, and the results compared with those obtained by others.

Theory of Radio Reflections from Electron-Ion Clouds—Von R. Eshleman

Approximations to the reflection coefficients of electron-ion clouds of various sizes, shapes, and densities are determined. A simplified approach to the problem is employed, wherein the total reflection from the low-density clouds is computed by summing the scattered wavelets from the individual electrons, while the high-density clouds are treated as total reflectors at the critical density radius. Some of the limitations of this method are discussed. While more elaborate determinations should be used in certain regions of cloud size and density, the method used here provides an overall, first order approximation of the effects of size, shape, and density on the reflecting properties of electron-ion clouds. Several possible applications of the theoretical results are outlined.

Discussion on Optimum Patterns for Endfire Arrays—R. L. Pritchard

The method of synthesizing an equal-minor-lobe directivity pattern suggested by Riblet for the broadside array was applied recently by DyHamel to the case of the endfire array. In the present paper, an alternative synthesis procedure is described, based directly on Dolph's method of synthesis for the broadside array. Advantages of this alternative procedure include (a) applicability to arrays having *even* as well as odd numbers of elements and (b) somewhat simpler equations for calculating relative currents for the elements of the array. This alternative procedure is described here in detail together with numerical examples for the seven-element array used by DuHamel and for a four-element array. A tabulation of equations for relative current amplitudes for arrays of 5, 7, 9, 11, and 13 elements is given in an Appendix. In conclusion, an alternative method of overdesigning a supergain antenna is described.

AUDIO

Vol. AU-3, No. 1, Jan.–Feb., 1955

D. W. Martin (Editorial)

PGA News

List of Chapter Chairmen, IRE Professional Group on Audio

Correction to "Equivalent Circuit Analysis of Mechano-acoustic Structures," B. B. Bauer

Equalization of Tone Controls on Phonograph Amplifiers—F. H. Slaymaker

The necessity for equalization in phonograph amplifiers is explained, and numerous examples are given. The difference between the action of equalization controls and tone controls is explained. Examples are shown in terms of response curves for a particular amplifier.

The Manufacture of High-Fidelity Magnetic Tape Records—R. C. Moyer

The equipment used in the manufacture of magnetic tape records by RCA is described. The choice of tape speed, track width, and preemphasis used is explained, and the desirability of standardization on these possible variables is emphasized.

Loudspeakers and Microphones—Leo L. Beranek

This paper covers the development of loudspeakers and microphones and gives many illustrations of commercially available units dating from 1915 to the present time.

BROADCAST AND TELEVISION RECEIVERS

Vol. BTR-1, No. 1, January, 1955

A Simple Method of Phase Compensation of Video Delay Lines—D. A. Gillen

A new method of phase compensation of delay lines of the distributed constant type is described in detail. This method produces a simple, compact line having superior phase and response characteristics. The phase compensation is effected by using inductive coupling within the line in the proper amount and phase

to achieve the desired correction. Such coupling does not introduce additional losses into the line and therefore the frequency response of the line, is not impaired by the introduction of the compensating means. Steady state and transient response measurements on various types of lines made using the above principle are shown and a practical method of designing a line to meet a given set of conditions is described.

Optimum Crystal Mixer Operation—The 1N82 Crystal—S. Deutsch

The dc voltage-current characteristic and the 10 mc/sec. noise temperature ratio of a group of 1N82 crystals were measured. The data thus obtained make it possible to calculate the condition for optimum mixer operation of the 1N82 crystal.

The most unpredictable element in the design of a television UHF tuner employing a crystal mixer is the image-frequency impedance, Z_i. Various quantities are accordingly plotted, as a function of Z_i, as the latter goes from zero to infinity in the resistance domain. For example, for an average 1N82 crystal, the optimum rf source resistance ranges from 397 to 801 ohms, the minimum conversion loss ranges from 3.32 db to 4.71 db, the optimum mismatch ranges from 2.02 to 2.74, and the IF output resistance ranges from 397 to 2031 ohms. The latter factor has proven to be extremely troublesome in television uhf tuners.

The average 1N82 noise temperature ratio, at 10 mc/sec, was found to be 4.9. This leads to a mixer noise temperature ratio, as a function of Z_i, of from 1.39 to 1.84 at 30 mc/sec. Finally, a set of curves is shown giving the optimum intermediate frequency and minimum mixer—IF amplifier noise figure for various 30 mc/sec. IF noise figures.

The method of analysis used here is applicable to any diode mixer. The minimum mixer—IF amplifier noise figure values will indicate the degree of success of any given 1N82 tuner design. In addition, the curves show that certain mixer circuit configurations are superior to others.

A New Self Oscillating Frequency Converter—D. E. Sunstein

A frequency converter utilizing the autodyne principle has been developed which has a feature not normally expected of the autodyne, namely, the ability to accommodate an automatic gain control signal. This ability is achieved by causing the amplitude of the oscillations to be controlled by an element external to the non-linear device doing the frequency converting. Conversion gain, noise performance, image rejection, as well as oscillator radiation, are generally equivalent to or better than that of the standard pentagrid circuits at broadcast band frequency.

Color Purity in Ungated Sequential Displays—G. S. Ley

(*Paper given at Radio Fall Meeting, Syracuse, New York, October 18, 1954*)

Color purity of ungated sequential displays involving the modification of the NTSC signal and the addition of harmonics is discussed. It is shown that in an idealized case of a continuous color sequence display very good color accuracy can be obtained by the addition of only the second harmonic of the color writing frequency. It is compared with an idealized reversing color sequence display using second harmonic only and another including fourth harmonic.

A block diagram of a receiver to give the continuous color sequence is given.

The Practical Aspects of the Color Subcarrier Synchronization Problem—Wolf J. Gruen

This paper deals with the behavior of different color subcarrier synchronizing circuits. These circuits include the simple AFC circuit, the hunting AFC circuit, the narrow-band—wide-band AFC circuit, and the crystal filter followed by balanced diode synchronous detectors. An evaluation of these circuits with respect to reliability of performance and simplicity of operation is presented.

The Second Detector—A Determinant of Fringe-Area Performance—J. E. Bridges

The linear diode detection process is shown to have an important influence on the performance of AM and television receivers in the fringe area. The diode detection process can seemingly increase selectivity; can, apparently, overload; and can, in effect, generate noise. Principal factors are the carrier-to-noise ratio, IF amplifier response, location of the carrier within the IF response, frequency of the modulating signal, and the degree of modulation. Special detection circuits which minimize undesired by-products of the detection process are capable of small improvements in the performance of television receivers in the fringe area.

A UHF-VHF Tuner Using Pencil Tubes—W. A. Harris and J. J. Thompson

The use of pencil tubes in a television tuner covering all channels in the uhf and vhf bands is discussed. The input and output resonant frequencies of the pencil tubes are high enough to allow the use of shunt-tuned circuits in a uhf-vhf tuner, and the feedback inductance is low enough to insure stability in the required frequency ranges. The shunt-tuning method has the advantages of providing small over-all dimensions and a moderate number of switching components.

The paper includes discussion of the measurements of the tube characteristics at high frequencies and the use of such data in the design of the tuner circuits.

ELECTRON DEVICES

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For contents and abstracts of papers, see Proc. I.R.E., pp. 1577–1580, October, 1954.

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

U.D.C. CHANGES

In anticipation of a new edition of the Universal Decimal Classification Abridged English Edition (BS 1000 A), certain changes in U.D.C. numbers will be made in this and subsequent issues. The new numbers used will be:

Radio astronomy: 523.16
Ultrasonics: 534 subdivisions with the special analytical subdivision -8 attached.
Sound recording and reproducing: 534.85
Electroacoustic problems, transduction, etc.: 534.86.

The Index to the Abstracts and References published in the PROC. I.R.E. from February 1954 through January 1955 is published by PROC. I.R.E. as Part 2 of the April, 1955 issue. It is also published by *Wireless Engineer* and included in the March 1955 issue of that journal, which may be purchased for $1 (including postage) from the Institute of Radio Engineers, 1 East 79th Street, New York 21, N. Y. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534.121.1 **311**
Excitation of Flexural Vibrations in Metallic Plates using Polycrystalline Barium Titanate—I. S. Zheludev. (*Zh. tekh. Fiz.*, vol. 24, pp. 1467–1473; August, 1954.) Production of Chladni's figures with square Fe, Al and duralumin plates is described. The $BaTiO_3$ powder is fixed in a given pattern in an adhesive between the plate and appropriately shaped foil electrodes, and is then polarized. The observed frequencies of vibration are in good agreement with values calculated from the Ritz formula; frequencies between 450 cps and 20.3 kc were produced.

534.213.4 **312**
A New Method of Computing Acoustical Filters—W. K. R. Lippert. (*Acustica*, vol. 4, no. 4, pp. 411–420; 1954.) Filters consisting of n equal sections, with nonreflecting output duct, are considered. Each section consists of an input and output duct of uniform cross section with a discontinuity between them. For loss-free filters, the characteristic transmission and reflection factors may be expressed in terms of B and β, where the transmission factor for a single section is $T = Be^{j\beta}$. Graphs are drawn for the corresponding quantities for 2, 3 and 4 sections. Measured and computed values for a filter with two rectangular bends, i.e. with two sections, are in satisfactory agreement. The characteristic factors for a filter with a reflecting output duct or with dissimilar sections may be derived from the fundamental case.

534.232:538.652 **313**
Some Remarks on the Analysis of Magnetostrictive Transducers—H. Nødtvedt. (*Acustica*, vol. 4, no. 4, pp. 432–438; 1954.) From the energy conditions in the system, an expression for the coupling coefficient is derived and used in plotting a curve of coupling-coefficient/polarizing-field for a Ni rod. Equivalent circuits of the rod are considered, and the analysis of either the admittance or impedance diagrams is shown to give the transducer operational parameters.

534.24–14 **314**
Radiation Pressure in Fluids—J. Mercier. (*Acustica*, vol. 4, no. 4, pp. 441–446; 1954. In French.) Radiation pressure is treated as a vector quantity equal to the product of the excess pressure and the velocity of displacement of a section of a plane wave. In the case of reflection from an obstacle, it must be calculated separately for incident and reflected waves. The cases of plane waves in an absorbent medium and spherical waves are also considered.

534.833.4 **315**
Sound Transmission through Spherical Bosses—A. Bergassoli, F. Canac and T. Vogel. (*Acustica*, vol. 4, no. 4, pp. 403–405; 1954. In French.) The frequency response curves for sound transmitted through (a) a hollow spherical boss, and (b) a disk having the same base dimensions were obtained. One plexiglass and three Al bosses were used, with different radii of curvature and/or heights. Sound insulation is considerably improved at frequencies below 1,000 cps by using a boss. Simple theory can account fairly well for these results.

534.844 **316**
The Acceptability of Artificial Echoes with Reverberant Speech and Music—A. F. B. Nickson, R. W. Muncey and P. Dubout. (*Acustica*, vol. 4, no. 4, pp. 447–450; 1954.) An artificial echo of adjustable intensity level and delay was added to speech and music signals. The subjectively determined acceptable-echo-level/delay relation is dependent on (a) the Haas effect (*Acustica*, vol. 1, pp. 49–58; 1951) for delay times up to 50 ms, (b) natural echoes in the recording room for the next few hundred milliseconds, and (c) the dynamic range of the program material for longer delay times.

534.86 **317**
Facilities for the Sound Recording and Observation of Interviews—G. F. Mahl, J. Dollard and F. C. Redlich. (*Science*, vol. 120, pp. 235–239; August 13, 1954.) The special conditions to be considered in relation to the recording of psychotherapeutic interviews are discussed. An installation at Yale University comprises a suite of two interviewing rooms with individual control rooms, and a waiting room forming a sound lock. Details are given of the low-noise air-conditioning system. A diaphragm-blanket acoustic treatment renders the sound natural although the rooms are quiet.

621.395.623.7 **318**
Loudspeakers and Enclosures—(*Audio*, vol. 38, pp. 27–28, 55; August, 1954.) A comprehensive survey, in simple terms, of the construction and operation of modern loudspeakers, including baffles and combination systems.

621.395.623.743 **319**
Electrostatic Speaker accents High Frequencies—M. Hobbs. (*Electronics*, vol. 27, pp. 143–145; November, 1954.) A loudspeaker for the range 7–15 kc comprises a stiff curved perforated copper-backed plate mounted rigidly with a gold-sprayed styroflex or polyethylene foil stretched over it, the capacitor thus constituted has a capacitance of about 4mμf. A polarizing voltage of the order of 250 v is required. The arrangement provides an economical design for radio receivers, used in conjunction with a cone loudspeaker for the lower frequencies.

621.395.625.3 **320**
New Portable Magnetic Recorder Type R85 for Commentaries—K. E. Gondesen. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 6, nos. 5/6, pp. 127–132; 1954.) The electrically driven recorder, amplifier unit and batteries are combined in a suitcase model weighing about 10 kg. The tape runs at 19 cm and reverses mechanically. One amplifier deals with recording and erasing, a separate one with reproduction and modulation control. A spool runs for 30 minutes and the accumulator which drives the motor permits continuous operation for four hours.

ANTENNAS AND TRANSMISSION LINES

621.315.212:621.372.51:621.396.67 **321**
New Antenna Coupler for U.H.F. Circuits—R. I. Stainbrook. (*Tele-Tech*, vol. 13, pp. 86–88, 181; August, 1954.) Coupling units for the 230–390-mc frequency band comprising cutaway coaxial-line sections are described. These couplers, when arranged in groups of 2–6 units, provide a system in which 2–6 transmitters, re-

ceivers and/or transmitter-receivers can operate simultaneously on a single antenna. The insertion loss for each one of a 6-unit group is 0.66 db.

621.372.2+621.372.8 322

The Transformation Properties of Lossless Six-Terminal Networks between Homogeneous Transmission Lines, their Representation by a Six-Terminal-Network Surface and their Significance in Measurement Technique—H. Lueg. (*Arch. elekt. Übertragung*, vol. 8, pp. 331–340; August, 1954.) The transfer characteristics of a six-terminal network representing a waveguide junction are derived by a matrix treatment and a three-dimensional surface is constructed, projection of which on to an appropriate plane gives the position of a voltage node on the input line as a function of the location of short circuits on the other two lines. Using the double-short-circuit technique [2850 of 1954 (Weissfloch)] input impedance can be determined directly by plotting a single curve. Power distribution between the outputs is dealt with and a method of determining parameter values for the equivalent circuits is illustrated.

621.372.2 323

The "Unit" Treatment of Impedance Irregularities and its Application to Long Lines—A. Rosen. (*Proc. IEE*, part IV, vol. 101, pp. 271–289; August, 1954. Digest, *ibid.*, part III, vol. 101, pp. 349–356; September, 1954.) Line irregularities are studied by a statistical method alternative to that based on correlation theory. It is assumed that the irregularities are built up of units each of which has a characteristic form; this may be produced by one independent physical factor operating over its own characteristic distance, but the amplitudes of the resulting unit irregularities occurring at different places along the line may be different, each being independent of any other. The length of the unit irregularity is assumed to be small compared with the length of the line, so that the attenuation over this length can be neglected. The procedure is to assemble a family of lines in which the amplitudes of the units are distributed at random and in which the units can occupy all positions with equal probability; the mean and mean-square values of the reflection and transmitted-echo coefficients for this family are calculated. The method is used to examine the effects of line irregularities on wide-band telephony and television transmissions. Results indicate that while the smoothness of the input-impedance/frequency characteristic is a suitable criterion for the regularity of coaxial-cable sections for telephony, this should be supplemented by pulse-reflection observations if the cable is intended for television. See also 1295 and 1668 of 1954.

621.372.2 324

Lower Modes of a Concentric Line having a Helical Inner Conductor—L. Stark. (*Jour. Appl. Phys.*, vol. 25, pp. 1155–1162; September, 1954.) Analysis is presented using cylindrical coordinates, the helix being assumed for convenience to have the form of an infinitely thin tape. Closed expressions are derived for the propagation factors of some of the lower modes. Numerical results are presented for a system with typical geometry.

621.372.2 325

Directional Electromagnetic Couplers—B. M. Oliver. (PROC. I.R.E., vol. 42, pp. 1686–1692; November, 1954.) General theory is given for the directional coupling action of ordinary transmission lines. Simple transmission-line couplers for the 1-mc-1-kmc band are discussed; the transmission characteristic is controlled by varying the coupling along the lines. An arrangement of four wires, or two wires and ground, is described which can serve simultaneously as directional coupler, filter and transformer.

621.372.2+538.566]:621.3.012 326

New Chart for the Solution of Transmission-Line and Polarization Problems—G. A. Deschamps. (*Elec. Commun.*, vol. 31, p. 188; September, 1954.) Addendum to 965 of 1954, acknowledging earlier work by Steiner (1361 of 1947).

621.372.2.029.6 327

Contribution on the Physics of the Multilayer Transmission Line—W. Wild, H. Larsen and W. Ebrel. (*Arch. elekt. Übertragung*, vol. 8, pp. 346–352; August, 1954.) A laminated coaxial line with five conducting layers is treated as four elementary pairs each of which is coupled to its neighbors by the impedance of the common conducting lamina. Taking account of this coupling, and assuming a uniform propagation coefficient, analysis shows that (a) resistive losses are low, since each conductor carries only the difference of the transmission currents of two adjacent pairs; (b) towards the center of the stack of laminae these transmission currents increase steadily while the conductor currents tend to zero. This result is generalized to apply to any number of layers; the coupling impedance greatly affects the magnitude and frequency characteristic of the line attenuation. Methods of coupling a laminated line to a two-wire system are briefly considered.

621.372.2.029.6 328

Input and Output Couplings for Coaxial Laminated Transmission Lines—H. Kaden and H. E. Martin. (*Arch. elekt. Übertragung*, vol. 8, pp. 387–403; September, 1954.) An examination is made of the energy loss due to mode conversion when a laminated line is excited by coupling it to a coaxial line. For a laminated line without metal core this loss is reduced to a minimum value of 0.064 N by making the radius of the inner conductor of the coaxial line 26.8 per cent and the inner radius of the outer conductor of the coaxial line 91.7 per cent of the radius of the laminated line. For a laminated line with metal core and radius ratio 3.65 the corresponding values are 0.0503 N, 39.8 per cent and 92.7 per cent. If in addition losses due to reflections at the junctions are eliminated by inserting transforming sections, total coupling loss at sending and receiving ends can be reduced to 0.128 N and 0.106 N respectively. The mode conversion losses can be further reduced by applying different potentials to the laminations; the design of networks for supplying these potentials is discussed.

621.372.8 329

Reciprocity Theorem applied to Waveguide Junctions with Elliptical Polarization—G. Pircher. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 694–696; September 20, 1954.] Matrix analysis is presented based on consideration of the combined effects, at a reference plane in each waveguide arm, of the waves traveling towards and reflected from the junction, each of which can be represented either by a pair of orthogonal linearly polarized components, or by a pair of components with opposite directions of rotation.

621.372.8 330

An Estimation of the Discontinuities introduced by Flange Couplings in Rectangular Waveguides—R. B. Nicholls. (*Elliott Jour.*, vol. 2, pp. 140–143; August, 1954.) From measurements made with couplings displaced abnormally sideways and with relatively large gaps left between flanges, theoretical predictions can be checked and the effects of smaller discontinuities estimated. Transverse displacements and twists occurring in flanged guides made to present dimensional tolerances are found to be of minor significance, but even small gaps (e.g. 0.001 inch at 3 cm λ) may give rise to serious reflections.

621.372.8 331

Propagation of Electromagnetic Waves over Plane Surface with Anisotropic Homogeneous Boundary Conditions—M. A. Miller [*Compt Rend. Acad. Sci.* (*URSS*), vol. 87, pp. 571–574; December 1, 1952. In Russian.] The surfaces considered are corrugated waveguides consisting of an assembly of thin parallel metal strips mounted on a metal base with or without guiding walls, as described by Rotman (33 of 1952). The dependence of the propagation modes on the ratio of surface reactance components is discussed.

621.372.8:621.318.134 332

Propagation of Transverse Electric (TE) Waves in a Waveguide containing Ferrites—A. Chevalier and E. Polacco. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 692–693, September 20, 1954.] Analysis is presented for the TE_{mo} mode propagated in a rectangular waveguide with an applied magnetic field perpendicular to the direction of propagation and parallel to the guide thickness.

621.396.67:621.372.8 333

Application of Homogeneous Boundary Conditions in the Theory of Thin Aerials—M. A. Miller. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1483–1495; August, 1954.) Theory of thin antennas and of slot radiators is applied to the case of antennas with given homogeneous anisotropic boundary conditions of the type $E_z+Z_1H_{\bar{\iota}}|_\Sigma$, $E_{\bar{\iota}}=-Z_2H_z|_\Sigma$ where Z_1 and Z_2 are different complex impedances which can be considered as components of the tensor of surface impedance Z; z and $\bar{\iota}$ are orthogonal coordinates on the surface of the antenna Σ. Such boundary conditions can be formulated for the interface of two media if the structure of the field in one medium is known and is independent of the structure of the field in the other medium (331 above). The formulas derived are applied to the case of a disk-loaded cylindrical waveguide-section, the dimensions of the disks being small compared with λ, and to a rectangular corrugated waveguide element in the form of a short trench.

621.396.673 334

The Characteristics of a Vertical Antenna with a Radial Conductor Ground System—J. R. Wait and W. A. Pope. (*Appl. Sci. Res.*, vol. B4, no. 3, pp. 177–195; 1954.) An approximate method is presented for calculating the input impedance; graphical methods are used in evaluating integrals involved in the solution. Results are plotted to show the dependence of the impedance on the number and length of the radial conductors for given frequency, antenna height and ground conductivity. The presence of the ground system modifies the radiated fields by only a few parts in a hundred under usual conditions. Calculated values for a 250-foot mast are in good agreement with experimental results for operation at 100 kc.

621.396.676:621.396.93 335

Ultra-High-Frequency Omnidirectional Antenna Systems for Large Aircraft—W. Sichak and J. J. Nail. (*Elec. Commun.*, vol. 31, pp. 202–213; September, 1954.) Reprint. See 2568 of 1954.

621.396.677 336

On the Optimum Illumination Taper for the Objective of a Microwave Aerial—J. W. Crompton. (*Proc. IEE*, part III, vol. 101, pp. 371–382; November, 1954.) "The Fourier-transform description of the radiation pattern from an aperture is used to obtain curves relating to over-all gain of an idealized antenna system, comprising an objective and waveguide flare, with the size and illumination of the aperture of the latter. For rectangular objectives the dimensions of the aperture of the flare may be varied independently in the two planes of symmetry and the over-all gain obtained by multiplying together the gain factors for the distributions across the objective in these two planes. In the case of the circular

objective this no longer applies, but curves are drawn showing the variation of the gain of the system as the size of a symmetrical feed is varied by equal amounts in the two planes of symmetry. It is shown that maximum over-all gain is obtained with a rectangular objective when the size of the feed is adjusted to give a primary pattern such that the illumination intensity at the edges of the objective is $8\frac{1}{2}$ db below that at the center. The corresponding optimum taper for circular objectives is $11\frac{1}{2}$ db. Primary radiation-patterns are drawn for 'constant,' 'cosine' and 'double-cosine' illuminations of the feed aperture. The variations of secondary beam-widths and first side-lobe levels with feed size for rectangular and circular objectives are discussed."

621.396.677.32 337
Resonance in a System of Coupled Vibrators and Tuning of "Wave-Channel" Aerials—D. M. Vysokovski. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 97, pp. 659–662; August 1, 1954. In Russian.] The calculation of the mean length of the director elements of this Yagi-type antenna to give maximum directivity is considered for the case of weak coupling between the elements, which are spaced at $\lambda/3$. The useful range of element lengths is from about 0.46 λ, which is optimum for one director, to about 0.41 λ, which is optimum for 20.

621.396.677.71.012.12 338
Polar Diagrams of Plane Slot Aerials—G. N. Kotschershewski. (*NachrTech.*, vol. 4, pp. 272–274; June, 1954.) A vertical $\lambda/2$ slot in a plane screen with dimensions of the order of one wavelength is considered as a magnetic dipole and an approximate expression is derived for the current density distribution in the horizontal plane. Polar diagrams are constructed for the equatorial horizontal plane for radiation (a) on one side only and (b) on both sides of the slot; these agree closely with experimental results.

621.396.677.833.2:535.42:538.566 339
On the Radiation Pattern of a Paraboloid of Revolution—J. P. Schouten and B. J. Beukelman. (*Appl. Sci. Res.*, vol. B4, no. 3, pp. 137–150; 1954.) The pattern is determined using the form of Huyghens' principle in which the field intensities in a given domain are expressed in terms of their tangential components on the boundary. Expressions involving infinite series of Bessel functions are obtained for the intensities of the distant field when an electric dipole is placed at the focus.

AUTOMATIC COMPUTERS

681.142 340
A New German Differential Analyser for Differential Equations—A. Walther. (*Z. Ver. Dtsch. Ing.*, vol. 96, pp. 755–758; August 1, 1954.) A mechanical instrument built for the Institute of Applied Mathematics, University of Bonn, and comprising 8 integrators.

681.142 341
A Curve Analyzer and General-Purpose Graphical Computer—C. S. French, G. H. Towner, D. R. Bellis, R. M. Cook, W. R. Fair and W. W. Holt. (*Rev. Sci. Instr.*, vol. 25, pp. 765–775; August, 1954.) The analog computer described comprises five photoelectric curve followers, two integrators, nine adding amplifiers and a pen recorder. The independent variable is represented by a low alternating voltage increasing linearly with time; the form of the equation is determined by the connections between the units, and the values of constants are set by potentiometers.

681.142 342
Harmonic Cancellation from Computing Voltage Source in Servo Analogue Computers—J. Alman. [*Elec. Eng.*, (N.Y.), vol. 73, pp. 711–713; August, 1954.] Analysis showing how harmonics of the carrier can be canceled out in the summing network of the servo amplifier by injecting appropriate harmonic signals into the quadrature rejection circuit. The analysis is valid for very small phase shift. The system is advantageous when harmonics cannot be sufficiently attentuated by filters.

681.142 343
Some Comparisons between Analogue and Digital Computers—W. E. Scott and A. C. D. Haley. (*Jour. Brit. IRE*, vol. 14, pp. 476–486; October, 1954.) The uses of analog and digital computers in industry are discussed and a large-scale machine of each class is described briefly.

681.142 344
Electronic Computers and Industrial Mathematics—B. D. Dagnall and R. L. Michaelson. (*Jour. Brit. IRE*, vol. 14, pp. 487–496; October, 1954.) The mode of operation of a high-speed digital computer is described in functional terms; programming procedure is illustrated in relation to typical applications.

681.142 345
The Application of Electronic Digital Calculating Methods to Punched Card Machines—J. H. Lucas. (*Jour. Brit. IRE*, vol. 14, pp. 497–503; October, 1954.) Principles of operation and design of the "Electronic Multiplying Punch" are discussed. An outline is given of the calculating process used for sterling and for decimal systems.

681.142:621.314.63 346
Welded Joints on Diodes reduce Computer Bulk—S. G. Lutz. (*Electronics*, vol. 27, pp. 154–158; November, 1954.) Diode-mounting methods are discussed taking up the least possible space while affording adequate accessibility. By using spot welding, the diode leads can be shortened to $<1/4$ inch. An elongated matrix and a honeycomb arrangement accommodating up to 128 diodes/in^3 are among the alternatives described.

681.142:621.385.012 347
Making Use of Curved Characteristics for the Higher Types of Calculation in Electronic Computers—H. Harmuth. (*Acta Phys. austriaca*, vol. 8, pp. 332–337; July, 1954.) Where errors up to a few per cent are permissible, deliberate use can be made of the curvature of tube characteristics in analog computers. Circuits for multiplying, dividing, squaring and obtaining square roots are discussed.

681.142:621.385.832 348
The Design of an Electrodynamic Multiplier—E. M. Deeley. (*Proc. IEE*, part IV, vol. 101, pp. 187–191; August, 1954. Digest, *ibid.*, part II, vol. 101, pp. 344–345; June, 1954.) Development of a design described earlier [1977 of 1949 (Deeley and Mackay)] for use in a differential analyzer.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.3:621.316.5 349
The Suppression of Switching Transients by a Shunt *RC* Circuit—E. A. Finlay. (*Proc. IEE*, part IV, vol. 101, pp. 266–270; August, 1954. Digest, *ibid.*, part II, vol. 101, pp. 575–576; October, 1954.) A detailed analysis is presented of the phenomena occurring when apparatus including shunt capacitance is suddenly switched across a low-impedance source of power through a short transmission line; conditions for the transient voltage in the shunt *RC* damping circuit to be nonoscillatory are deduced. It is proved that the current through the switch can be made unidirectional. Numerical illustrations are given.

621.3.018.75:621.387.4 350
Fast Single-Channel Pulse Amplitude Analyser—W. L. Buys. (*Jour. Phys. Radium*, vol. 15, p. 582; July/September, 1954.) Designed for use with proportional or scintillation counters, the equipment has a channel-edge position stable to within +0.02v under operation for periods up to 10 hours.

621.3.018.75:621.387.4 351
Description of a Recording Instrument—P. Desneiges and A. Pages. (*Onde élect.*, vol. 34, pp. 700–704; August/September, 1954.) The recorder was designed for operation with the pulse analyzer described by Guillon (39 of February). It has 25 channels each comprising a decade counter and em register; one unit is recorded in the *n*th channel for every train of *n* pulses. Details are given of circuits and operation.

621.3.066.6 352
Migration of Material in Silver and Platinum Contacts—E. Holm and R. Holm. (*Z. angew. Phys.*, vol. 6, pp. 352–361; August, 1954.) An experimental investigation of deformation caused exclusively by the passage of current.

621.314.2 353
Component Design Trends—Iron-Core Transformers run Smaller and Hotter—F. Rockett. (*Electronics*, vol. 27, pp. 136–142; November, 1954.) A survey of the developments in core materials, coatings and insulating materials and of new construction methods which have in combination made possible a 50 per cent reduction of size of iron-core components over a period of about ten years while permitting operation at much higher temperatures.

621.314.7:[621.373.5+621.375.4 354
Surface-Barrier-Transistor Measurements and Applications—R. J. Turner. (*Tele-Tech.*, vol. 13, pp. 78–80, 165; August, 1954.) Circuits for determining the transistor characteristics are described and examples of applications taking advantage of the high frequency cut-off of the surface-barrier transistor are given. These include a two-stage video amplifier of bandwidth 3.2 mc and gain 28 db, a bandpass amplifier for 5–55 mc with measured power gain of 12.2 db at 55 mc, a crystal-controlled oscillator delivering 400 μw at 50 mc with an efficiency of 25 per cent, a mixer with a conversion gain, from 55 mc to 5 mc, of 8.6 db, and a nonsaturating flip-flop circuit with rise time 0.07 μs and fall time 0.15 μs.

621.316.8:[621.317.353+621.396.822 355
New Resistor Voltage-Coefficient Tester—L. A. Rosenthal and A. S. Louis. (*Tele-Tech.*, vol. 13, pp. 62–64, 114; September, 1954.) The variation of resistance with applied voltage is determined from measurements of the third harmonic generated by the resistor on application of a sinusoidal voltage. The basic circuit is described and some experimental results are given. The measurements of resistor noise at af is also briefly described.

621.316.8.029.5 356
The Design of a Radio-Frequency Coaxial Resistor—C. T. Kohn. (*Proc. IEE*, part IV, vol. 101, p. 299; August, 1954.) Discussion on 2019 of 1954.

621.316.84:621.372.221 357
On the Use of Commercial Wire-Wound Resistors in High-Level Video Amplifiers—A. J. Seyler. [*Proc. IRE* (Australia), vol. 15, pp. 219–227; September, 1954.] By compensating the inherent inductance by means of a parallel *RC* network, commercial wire-wound resistors can be adapted for use in high-power video amplifiers. Practical arrangements are illustrated and results of measurements are compared with calculated values.

621.316.86+621.316.89]:621.396.822 358
Measurements on Current Noise in Carbon Resistors and in Thermistors—K. M. vanVliet, C. J. van Leeuwen, J. Blok and C. Ris. (*Physica*, vol. 20, pp. 481–496; August, 1954.) Measurements are reported on carbon and similar resistors in the frequency range from 2 cps to

1.2 mc, and on negative-temperature-coefficient thermistors between 2 cps and 50 kc. In carbon resistors the current noise is proportional to the square of the current (i) and inversely proportional to approximately the first power of frequency (ν). In thermistors the noise level is lower and the spectrum shows deviations from the $1/\nu$ relation. The spectral density in the part of the spectrum where the $1/\nu$ relation holds is proportional to $i^{1.2}$. This is in agreement with measurements by Brophy (*Jour. Appl. Phys.*, vol. 25, pp. 222–224; February, 1954.)

621.318.42 359

Calculation of Iron-Cored Inductances for Direct Current—P. Ross. (*Rev. gén. élect.*, vol. 63, pp. 551–559; September, 1954.) A method described by Kammerloher (29 of 1949) is used. The position of the maximum is determined corresponding to the optimum gap length for laminations of given shape and size; variation of the maximum with number of ampere-turns is investigated. Examples are worked out.

621.319.46 360

A Fixed Gas-Dielectric Capacitor of High Stability—W. K. Clothier. (*Proc. IEE*, part II, vol. 101, pp. 453–459; August, 1954.) Factors contributing to instability of capacitors are reviewed and design principles are proposed for parallel-plate types suitable for use under laboratory conditions. A 1,000-$\mu\mu$f capacitor embodying these principles is described; the performance figures indicate that a considerable gain in stability can be achieved without serious complication in design.

621.37/.39.029.4:004.1 361

Better Audio Specs Needed—N. Grossman. (*Audio*, vol. 38, pp. 36, 38; September, 1954.) The significance is explained of technical terms commonly used in catalogue descriptions of af amplifiers, loudspeakers and af output transformers and a list is given of additional information required for a full technical description of components and apparatus.

621.37/.39].029.63/.64 362

Microstrip Kit—(*Elec. Commun.*, vol. 31, p. 214; September 1954.) Note of a commercially available set of microstrip components permitting rapid construction of circuits for operation in the frequency range 1–8 kmc.

621.372 363

A Mathematical Technique for the Analysis of Linear Systems—J. R. Ragazzini and A. R. Bergen. (PROC. I.R.E., vol. 42, pp. 1645–1651; November, 1954.) The technique of the z transformation developed for investigating sampled-data systems [1801 of 1953 (Ragazzini and Zadeh)] is applied to the numerical solution of continuous linear systems. A model of the linear system is devised in which sampling is introduced at some convenient point. A polygonal approximation to the sampled function is reconstructed by means of a holding operator, and the output of the system is computed as a train of pulses.

621.372 364

Properties of some Wide-Band Phase-Splitting Networks—M. Dishal. (PROC. I.R.E., vol. 42, p. 1698; November, 1954.) Errors in the design equations presented by Luck (1614 of 1949) are indicated and alternative methods of deriving the design parameters are mentioned.

621.372 365

The Return-Difference Matrix in Linear Networks—L. Tasny-Tschiassny. (*Proc. IEE*, part IV, vol. 101, pp. 299–300; August, 1954.) Discussion on 662 of 1954.

621.372 366

The Solution of a Certain Differential Equation—E. B. Moullin. (*Proc. IEE*, part IV, vol. 101, pp. 290–299; August, 1954. Digest, *ibid.*, part I, vol. 101, pp. 204–206; July, 1954.) A method of obtaining approximate solutions to a class of nonlinear network problems is discussed. The method was developed in relation to the problem of a capacitor in series with a nonlinear resistor of the SiC type, and its application is illustrated by reference to this problem, the resistor characteristic being assumed cubic and the applied voltage simple harmonic. An exact steady-state solution for this network is unobtainable, but an approximate solution is obtained by choosing the constants in the formula for the voltage across the resistor such that the applied voltage has no third-harmonic component.

621.372:512.831 367

Matrix Analysis of Linear Time-Varying Circuits—L. A. Pipes. (*Jour. Appl. Phys.*, vol. 25, pp. 1179–1185; September, 1954.)

621.372.413 368

On Cavity Resonators with Nonhomogeneous Media—A. Cunliffe, R. N. Gould and K. D. Hall. (*Proc. IEE*, part IV, vol. 101, pp. 215–218; August, 1954. Digest, *ibid.*, part III, vol. 101, pp. 192–193; May, 1954.) "Perturbation theory is used to find the frequencies of oscillation and Q-factors of a cavity containing a medium for which the permittivity and the permeability are arbitrary functions of position. The theory is applied to an H_{011} cylindrical cavity containing a disc of dielectric which does not extend over the complete cross section of the cavity."

621.372.413:530.145 369

Quantum Effects in the Interaction between Electrons and High-Frequency Fields: Part 1.—I. R. Senitzky. (*Phys. Rev.*, vol. 95, pp. 904–911; August 15, 1954.) The problem of determining at what frequency quantum-mechanical effects become apparent is approached by analyzing the velocities of electrons after passage through a cavity resonator. From the formulas derived a calculation is made of the ratio of (minimum possible mean-square deviation of velocity) to (velocity increment due to the field). Assuming the volume of the cavity to be of the same order as λ^3 and the initial electron velocity to be one-tenth that of light, this ratio becomes unity for a frequency of the order of 3×10^{12} cps.

621.372.5 370

Synthesis of Ladder Networks to give Butterworth or Chebyshev Response in the Pass Band—E. Green. (*Proc. IEE*, part IV, vol. 101, pp. 192–203; August, 1954. Digest, *ibid.*, part III, vol. 101, pp. 115–118; March, 1954.) General formulas are derived for the design of low-pass ladder networks. See also 1267 of 1953.

621.372.5 371

Synthesis of Constant-Time-Delay Ladder Networks using Bessel Polynomials—L. Storch. (PROC. I.R.E., vol. 42, pp. 1666–1675; November, 1954.) A low-pass minimum-phase transfer function is constructed which produces a maximally flat time-delay/frequency characteristic and a nearly Gaussian impulse response. Evaluation of delay and loss characteristics is simplified since the transfer function can be expressed in terms of tabulated functions. Synthesis of the corresponding ladder network using low-Q elements is described. Characteristics of a class of related nonminimum-phase transfer functions are also discussed.

621.372.5 372

Gain-Bandwidth Limitations on Equalizers and Matching Networks—H. J. Carlin. (PROC. I.R.E., vol. 42, pp. 1676–1685; November, 1954.) Comprehensive theory is presented and new results are given for the optimum voltage transfer to an arbitrary load.

621.372.5 373

Synthesis of Transfer Functions by Active *RC* Networks—D. B. Armstrong and F. M. Reza. (*Trans. I.R.E.*, vol. CT-1, pp. 8–17; June, 1954.) A criterion is presented for determining whether a prescribed transfer function can be synthesized by means of an active *RC* network with a single feedback loop. It is shown that any quadratic function can be synthesized by this means; the arbitrary function can be treated by breaking it down into quadratic elements. Alternative methods involving (a) inversion of the transfer function or (b) a ladder-network realization are also discussed.

621.372.5 374

Power Gain in Feedback Amplifier—S. J. Mason. (*Trans. I.R.E.*, vol. CT-1, pp. 20–25; June, 1954.) A search is made for a network figure of merit which remains invariant for different connection conditions, a quantity U is found satisfying this requirement and identified as the available power gain when the coupling is chosen to make the device unilateral. The quantity U may be useful for classifying active networks; its value is unity for a gyrator. The method directs interest to power gain and signal-flow considerations rather than to voltage gain and the network interpretation of an impedance matrix.

621.372.54 375

Determination of Energy and Phase Characteristics of Electrical Filters—G. N. Rapoport. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1496–1498; August, 1954.) The Leontovich theorem, which states that in nondissipative media the velocity of propagation of em energy is equal to the group velocity, is applied to the case of a filter, considered as an equivalent waveguide system. The velocity of propagation of energy can be calculated when the dispersion is known.

621.372.54.029.4:534.113 376

Tunable Audio Filters—G. Zelinger. (*Electronics*, vol. 27, pp. 173–175; November, 1954.) Electromechanical filters of clamped-resonant-reed type are described. The Q factor can be varied from 100 to 400 over the frequency range 60–700 cps.

621.372.543.029.3 377

L. F. Vibration Filters—F. M. Gol'tsman. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1350–1353; July, 1954.) The application of vibration galvanometers as lf narrow-band electrical filters is considered theoretically; experimental results for frequencies of 27–420 cps are tabulated and presented graphically. The corresponding bandwidths varied between 0.6 and 3.8 cps.

621.372.543.3:621.372.22 378

Inhomogeneous Transmission Lines as Filters—A. L. Feldstein. (*NachrTech.*, vol. 4, pp. 264–266; June, 1954.) Mathematical treatment of the filtering action of a transmission line with periodic structure. Design is based on an approximate integral formula expressing reflection coefficient as a function of characteristic impedance ρ. Its application in a simple case shows that the center frequency is determined by the length of a section, and the width of the stop band is dependent only on the total length, i.e. the number of sections. The formula gives results in good agreement with experiment for a single-section filter with ratio $\rho_{max}/\rho_{min}=8$.

621.372.6 379

Improved Matrix and Determinant Methods for Solving Networks—W. S. Percival. (*Proc. IEE*, part IV, vol. 101, pp. 258–265; August, 1954. Digest, *ibid.*, part III, vol. 101, pp. 278–279; July, 1954.) "It is shown that a linear electrical network, which may include tubes and transformers, can be represented topologically by a network composed entirely of elements of a single type. In its most general form an element of this type comprises four nodes and two directed branches. It is also

shown that a linear electrical network can be represented algebraically by a matrix, termed an H-matrix. If the elements of the network are shown in the network diagram in the form of elements of the above type, the H-matrix can be written down directly from the network diagram."

621.373.4 380
Effect of Electrical Fluctuations on Valve Oscillator—P. I. Kuznetsov, R. L. Stratonovich and V. I. Tikhonov. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 97, pp. 639–642; August 1, 1954. In Russian.] The effect of fluctuations which are slow compared with the oscillation frequency is considered theoretically and a method of calculating the effect is developed based on the Einstein-Fokker equation.

621.373.4:621.376.3 381
Frequency Control of Short-Wave Oscillators by means of an Electrodeless Gas Discharge—E. Häusler and B. Koch. (*Z. Naturf.*, vol. 9a, pp. 182–183; February, 1954.) Continuation of investigations reported previously [362 of 1954 (Koch)]. With lower oscillator power, higher gas pressure and looser feedback coupling a characteristic of quite different type is obtained, giving a rapid decrease of frequency with increase of anode voltage over a narrow range.

621.373.4:621.385.029.65 382
Possibility of obtaining Very Short Continuous Waves, using a Reflex Klystron supplying Energy at High-Order Harmonics of the Fundamental Oscillation—J. Bernier and H. Leboutet. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 796–798; October 4, 1954.] A formula is given for determining the wavelengths of the natural resonances of the klystron cavity. The dependence of these wavelengths on the grid separation is shown graphically, together with curves giving the values corresponding to exact harmonics of the fundamental. The two families of curves intersect at an infinite number of points, at each of which the cavity is capable of oscillating at both fundamental and harmonic frequency. Using a klystron with a beam voltage of about 2 kv and current 65 ma, and reflector voltage 1–1.5 kv, wavelengths down to 1.7 mm (24th harmonic of 4.08 cm) have been generated with a power of some tens of microwatts.

621.373.431.1 383
Method for Elimination of Effect of Stray Capacitances in the Production and Amplification of Step Voltages—W. Kroebel. (*Z. angew. Phys.*, vol. 6, pp. 293–297; July, 1954.) Discussion is presented of the charge and discharge mechanism of a multivibrator circuit; to obtain a steep leading pulse flank, a rapid discharge of the negative charge accumulated at one of the anodes is necessary. This is achieved by using the anode-dynode path of an auxiliary secondary-emission tube. Pulses of amplitude up to 400 v, with flanks of slope about 10^{10} v/second were obtained. Application of the principle in frequency dividers is also described.

621.375.1 384
General Properties of Electromagnetic Amplifiers—D. A. Bell. (*Wireless Eng.*, vol. 31, pp. 310–319; December, 1954.) The relations between gain, bandwidth or time constant, and power rating are discussed; results to be expected from use of feedback are examined. The amplifiers considered include thermionic-tube, magnetic and rotating-machine types. Devices for improving the shape of the pass-band normally involve delay; distributed amplifiers permit improvement of the gain-bandwidth parameter but not of the gain/time-constant ratio.

621.375.2.024:621.385.5 385
"Starved Amplifiers"—G. E. Kaufer. (*Electronic Eng.*, vol. 26, pp. 498–503; November, 1954.) Advantages of the starved amplifier (i.e. tube operated with very low anode and screen voltages) for high-gain amplification of dc are indicated. Relevant tube data must be determined by the designer; typical curves for Type-6SG7 and 6AG5 miniature pentodes are presented and discussed.

621.375.221 386
Design of Video Couplings—D. G. Sarma. (*Wireless Eng.*, vol. 31, pp. 327–334; December, 1954.) A method is presented for designing low rise-time video-amplifier coupling networks with three or four independent elements by the proper location of complex zeros and poles of the network function. The relations between rise-time, overshoot and delay are determined. Graphs are presented for determining the type of response of a circuit with given constants, or the circuit constants for a given type of response.

621.375.222 387
Zero Stabilization of Directly-Coupled Amplifiers—E. H. Frost-Smith and A. R. B. Churcher. (*Elliott Jour.*, vol. 2, pp. 136–139; August, 1954.) Details are given of circuits in which stability of a direct-coupled amplifier is achieved by use of a second amplifier in a feedback arrangement. The auxiliary amplifier is highly stable but has a low cut-off frequency; hence low-frequency signals are pre-amplified, giving enhanced signal/noise ratio at these frequencies. For high-frequency signals the auxiliary amplifier is by-passed.

621.375.3 388
An Even-Harmonic Magnetic Amplifier—P. D. Atkinson and A. V. Hemingway. (*Electronic Eng.*, vol. 26, pp. 482–485; November, 1954.) The circuit is an adaptation of the inverter described by Frost-Smith (73 of 1954); a typical figure for the power gain is 1,000, and for the output 1 mw. The zero drift is less than the shift produced by an input power of 10^{-11} w. The input impedance ranges up to about 3 kΩ, making it suitable for use with many types of transducer, but the poor response time restricts application to cases where the input varies slowly. Multistage amplifiers of this type are useful for control and measurement purposes.

621.375.3 389
Fast-Response Magnetic Amplifiers—G. E. Hughes and H. A. Miller. (*Trans. AIEE*, Part I, *Communication & Electronics*, vol. 73, pp. 69–75; March, 1954. Digest, *Elec. Eng.* (*N.Y.*), vol. 73, p. 723; August, 1954.) Expressions are given for the number of control turns and the gain of magnetic amplifiers of the type described by Ramey (3503 of 1953), in which load and control circuits operate on the magnetic state of the core on a time-sharing basis.

621.375.3:538.244 390
Effect of Circular Field on Form of Longitudinal Magnetization Curve—R. E. Ershov. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1508–1512; August, 1954.) The variation was investigated experimentally of the magnetization of an iron tube placed in a coaxial magnetic field combined with a circular field produced by a current flowing in a conductor placed along the axis of the tube. The results, which are presented graphically, are of interest in the construction of transformers for amplification of weak currents.

621.375.4 391
Feedback Simplifies Transistor Amplifiers—S. Schenkerman. (*Electronics*, vol. 27, pp. 129–131; November, 1954.) "Voltage amplification in transistor stages is obtained using a degenerative feedback path shunting the output and in series with the input. Use of interstage transformers or grounded-collector stages is avoided."

621.375.4.029.4 392
A Transistor Remote Amplifier—P. Penfield, Jr. (*Audio*, vol. 38, pp. 26–27; September, 1954.) A portable af pre-amplifier using Type-CK-721 point-contact transistors is described; precautions necessary to avoid damage to the transistors during construction of the equipment and during use are indicated.

621.375.5:537.227 393
Ferroelectrics and the Dielectric Amplifier—Mason and Wick. (See 447.)

621.375.9:537.312.8 394
Galvanomagnetic Low-Frequency Amplifier—E. Justi and H. J. Thuy. (*Z. Naturf.*, vol. 9a, pp. 183–184; February, 1954.) A simple arrangement is described based on the high magnetoresistance of Bi at low temperature. The device is much less noisy than an ordinary tube or transistor. See also 3179 of 1954 (Thuy).

621.376.239 395
Magnetic Demodulation with Sinusoidal Variation of the H. F. Induction Flux—G. Meinshausen. (*Arch. elekt. Übertragung*, vol. 8, pp. 373–386; September, 1954.) With magnetic demodulation using sinusoidal rf current in the demodulating reactor [3400 of 1953 (Pungs and Meinshausen)], the rf flux is distorted. By arranging instead for the rf flux to be sinusoidal, the power can be increased. Analysis is given based on analogy with diode theory, the flux being treated as the equivalent of the voltage across the diode. Essential differences between magnetic and diode demodulators are discussed.

GENERAL PHYSICS

53:519.2 396
Statistical Concepts in Theoretical Physics—R. Fürth. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 273–276; August, 1954.) A plea for the teaching of classical physics on an empirical basis on statistical lines, to make easier the transition to quantum physics.

530.112 397
Proposal for a New Aether Drift Experiment—H. L. Furth. [*Nature* (*London*) vol. 174, pp. 505–506; September 11, 1954.] Further elucidation of the proposal made previously (1726 of 1954).

535.23 398
The Maximum of the Planck Energy Spectrum—R. N. Bracewell. [*Nature* (*London*) vol. 174, pp. 563–564; September 18, 1954.] In the formula generally used for locating the maximum in the energy spectrum of a black body, the spectral distribution is based on the amount of energy lying between two closely spaced wavelengths. If the distribution is considered on a frequency basis instead of a wavelength basis, a very different value is obtained for the position of the maximum; for a body at 6,000 degrees (roughly the temperature of the sun) the maximum then falls at 8,497 Å, which is outside the visible region. Re-examination of Planck's formula leads to a Wien-type law yielding a natural maximum at 6,115 Å for the 6,000 degree spectrum; the location of this maximum does not alter on changing from a wavelength to a frequency basis.

535.33:538.56.029.6 399
Apparatus for Microwave Spectroscopy—M. W. P. Strandberg, H. R. Johnson and J. R. Eshbach. (*Rev. Sci. Instr.*, vol. 25, pp. 776–792; August, 1954.) The design of a spectrograph using Stark modulation and a crystal or bolometer detector is discussed, and a detailed description is given of a particular equipment.

535.42 400
Diffraction of Waves by a Wedge—F. Oberhettinger. (*Commun. pure Appl. Math.*, vol. 7, pp. 551–563; August, 1954.) Green's functions of the wave equation for a wedge are given in the form of an integral representation for the two-dimensional case, and are evaluated by the residue theorem. The half-plane is considered as the limiting case of a wedge with

angle 2π. The analysis is extended to the three-dimensional case.

535.42 401
A Uniqueness Theorem and a New Solution for Sommerfeld's and Other Diffraction Problems—A. S. Peters and J. J. Stoker. (*Commun. pure Appl. Math.*, vol. 7, pp. 565–585; August, 1954.)

535.42:538.566 402
On the Scalar Diffraction by a Circular Aperture in an Infinite Plane Screen—A. T. de Hoop. (*Appl. Sci. Res.*, vol. B4, pp. 151–160; 1954.) Application of variational principles reduces the problem of finding the aperture distribution to the solution of an infinite system of linear equations. An investigation is made of a particular system of equations based on a convenient expansion of the aperture distribution given by Bouwkamp.

535.42:538.566 403
General Approximation Method for solving Diffraction Problems—M. Jessel. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 239, pp. 753–756; September 27, 1954.] A system of integro-differential equations is established for the boundary values. A known variational principle is used for calculating certain functionals involved.

537.291+538.691 404
Motion of Charged Particles in a Homogeneous Magnetic Field, with Superposed Magnetic Field of Linear Current and Electric Field of Cylindrical Capacitor—V. M. Kel'man and S. Ya. Yavor. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1329–1332; July, 1954.) The problem considered theoretically by Kel'man and Rodnikova (75 of 1953) is extended to include the effect of a magnetic field parallel to the axis of the cylinder. Calculated results for several particular cases are presented graphically.

537.311.31:621.317.332.029.65 405
R. F. Conductivity in Copper at 8 mm Wavelengths—J. S. Thorp. (*Proc. IEE*, part III, vol. 101, pp. 357–359; November, 1954.) Measurements at 8 mm λ indicate that reduction of conductivity below the dc value is caused not only by surface roughness but also by the presence of surface layers having dc conductivity lower than that of the bulk material, and by stress in the bulk material. These effects can be overcome by etching and annealing, or by covering the surface layers by a process involving oxidation and reduction.

537.311.33:537.323 406
Limits of Practical Application of Pisarenko's Formula—T. A. Kontorova. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1291–1297; July, 1954.) Pisarenko's formula expresses thermoelectric power in terms of charge-carrier concentration and effective mass and of absolute temperature. A constant factor is included to take account of scattering of the charge carriers in the crystal; its value depends on the type of scattering and crystal lattice. Calculations show that the errors are small (<5 per cent) up to carrier concentrations of about 2×10^{19} in atomic-lattice and 7.2×10^{19} in ionic-lattice crystals at 300 degrees K. The errors are discussed, and compared with those of other formulas. Results are tabulated.

537.36 407
The Motion of Electricity—J. J. Gruetzmacher. (*Naturwiss.*, vol. 41, p. 368; August, 1954.) Preliminary experiments are described on the mechanical transportation of electric charge, e.g. by blowing streams of ionized air, or by causing such streams to flow in tubes under a pressure difference. *G-M* tubes are used as detectors, with loudspeaker indicators. Currents up to 10^{-14}a have been produced.

537.525:546.291 408
Energy Distribution Function of Electrons in Pure Helium—F. H. Reder and S. C. Brown. (*Phys. Rev.*, vol. 95, pp. 885–889; August 15, 1954.)

537.525:546.291 409
Microwave Determination of the Probability of Collision of Electrons in Helium—L. Gould and S. C. Brown. (*Phys. Rev.*, vol. 95, pp. 897–903; August 15, 1954.)

537.525.72:538.63 410
Some Effects in Inductively Coupled Electrodeless High-Frequency Gas Discharges with Superposed Magnetic Field—N. Neuert, H. J. Stuckenberg and H. P. Weidner. (*Z. angew. Phys.*, vol. 6, pp. 303–310; July, 1954.) The effects noted by Lindberg et al. (85 of 1953) were further investigated. The resonance effect appears to be due to electrons moving initially in circular orbits coaxial with the tube wall rapidly acquiring energy in the alternating field. An application of the effect in a hf ion source is briefly described. For another application see 90 of February (Koch and Neuert).

538.221 411
Extension of the Spin-Wave Theory of Ferromagnetism to Higher Temperatures—G. Heber. (*Z. Naturf.*, vol. 9a, pp. 91–97; February, 1954.)

538.3 412
Possibility of considering Potentials and Fields as Quantities having Density: New Type of Quadrivector—É. Durand. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 239, pp. 751–753; September 27, 1954.]

538.312 413
On Multipole Expansions in the Theory of Electromagnetic Radiation—C. J. Bouwkamp and H. B. G. Casimir. (*Physica*, vol. 20, pp. 539–554; August, 1954.) "A new method is developed for expanding the electromagnetic field of radiating charges and currents in multipole components. Outside a sphere enclosing all sources, the field is represented in terms of Debye potentials which are shown to be closely related to the radial components of the electric and magnetic vectors. Attention is drawn to remarkably simple source representations of these radial components. A discussion of vector-potential versus Debye-potential representation of multipole fields is included. Only familiarity with ordinary vector calculus is required."

538.56 414
The Radiation Green's Functions—J. G. Linhart. (*Jour. Frank. Inst.*, vol. 258, pp. 99–112; August, 1954.) Analysis for the radiation field of a group of charges is derived using functions similar to Green's function of potential theory but applied to the time-variables problems of radiation theory. The method presents a model of interaction between a quantum of movement of a unit point charge and the corresponding excitation of a set of eigen-fields; it is illustrated by treating the case of a point charge rotating in a cylindrical cavity.

538.566:538.614 415
Propagation of Electromagnetic Waves in a Gyrotropic Medium—M. A. Gintsburg. [*Compt. Rend. Acad. Sci. (URSS)*, vol. 97, p. 572; August 1, 1954, In Russian.] Correction to paper abstracted in 2927 of 1954.

538.632/.633 416
Direct Comparison of Hall Effect and Corbino Effect—W. J. Poppelbaum. (*Helv. Phys. Acta*, vol. 27, pp. 355–394; August 31, 1954. In French.) By using a cylindrical sample immersed in radial magnetic field, a transverse current is produced when an alternating voltage is applied between the ends of the cylinder; this transverse current gives rise to an induced voltage in a coaxial coil. Both the Corbino and the Hall effects can be measured with this arrangement, thus providing two independent means of determining the Hall constant; the discrepancy between the two results is <5 per cent.

621.3.032.44 417
The Distribution of Temperature along a Thin Rod Electrically Heated in Vacuo: Parts 2–4—S. C. Jain and K. S. Krishnan. (*Proc. Roy. Soc. A.*, vol. 225, pp. 1–32; August 6, 1954.) The theoretical solution presented in part 1 (2375 of 1954) is considered in the light of formulas discussed by Stead (*Jour. IEE*, vol. 58, p. 10; 1920) relating the temperature distribution in a finite rod to that in an infinitely long rod heated by the same current. A useful expression is found for the temperature at the center of a finite long rod as a function of its length. Detailed experimental verification of the main theoretical results is reported. Empirical formulas proposed by Worthing are shown to follow from the theory developed.

53.05 418
The Physics of Experimental Method [Book Review]—H. J. J. Braddick. Publishers: Chapman & Hall, London, Eng., 404 pp., 35s (*Phil. Mag.*, vol. 45, p. 884; August, 1954.) Sections are included on relevant mathematics, mechanical construction and materials, vacuum technique, electrical measurements and electronics, optics and photography, natural limits of measurements, and techniques of nuclear physics.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 419
A Comparison of the Intensity Distribution of Radio-Frequency Radiation with a Model of the Galactic System—G. Westerhout and J. H. Oort. (*Bull. astr. Insts Netherlds*, vol. 11, pp. 323–333; August 26, 1951. In English.) A comparison is made between the radiation to be expected from objects distributed like the common stars and the observed rf radiation. Satisfactory agreement with Bolton and Westfold's observations on 100 mc (348 of 1951) is obtained for the regions where the radiation is considerable. Near the galactic poles and in the hemisphere opposite the center the observed temperatures are about 600 degrees in excess of the calculated values; this excess is possibly due to a background of distant extragalactic nebulae.

523.16 420
Methods and Results of Radio Astronomy—H. Siedentopf. (*Z. angew. Phys.*, vol. 6, pp. 376–384 and 422–430; August and September, 1954.) A survey with over 100 references.

523.5:621.396.96 421
Radio Observations of Meteors—P. M. Millman. (*Science*, vol. 120, pp. 325–328; August 27, 1954.) A summarized general account, with literature references, of the most significant results obtained during the seven years up to the end of 1953.

523.746.5 422
Present Phase of Solar Cycle—D. E. Trotter and C. Pecker. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 239, pp. 633–635; September 6, 1954.] Observations of the intensity of the coronal 5303-Å green line suggest that the 1954 minimum of solar activity is flatter and lower than that of 1944; a first indication of the rise of the new cycle was observed at the beginning of August, 1954. The next maximum is expected to occur between summer, 1959 and the end of 1960 and to be rather low.

550.385 423
The Prediction of Geomagnetic Disturbances—J. C. Pecker and W. O. Roberts. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 239, pp. 635–637; September 6, 1954.] Shapley's method for predicting geomagnetic indexes (*Trans. Amer. Geophys. Union*, vol. 28, p. 715; 1947) is modified to take account of (a) the variation

of the periodicity of the disturbance about the 27-day average, and (b) the maxima observed in the values of the indexes in March and September.

551.510.53+621.396.11 424
Rocket Exploration of the Upper Atmosphere—R. L. F. Boyd, M. J. Seaton and H. S. W. Massey, Eds. (*Jour. Atmos. Terr. Phys.*, special supplement, vol. 1, p. 376; 1954.) The text is given of 49 papers read at a conference at Oxford in August, 1953. Sections 1–3 cover rocket techniques, atmospheric phenomena and composition of the high atmosphere. Section 4 deals with the ionosphere in particular and includes papers on wave-propagation measurements made using rockets (reported earlier), a specially designed antenna used on dry rocky soil for short-range sky-wave measurements at lf, an analysis of a method of determining charge densities from rocket-tracking data, and a suggested program of rocket-borne-magnetometer measurements. Section 6 covers laboratory studies, including an experiment on cross-modulation.

551.510.535 425
Study of Layer Splitting in the Ionosphere—O. Burkard. (*Arch. Met. A*, *Wien*, vol. 7, pp. 283–291; June 30, 1954.) An equation is derived expressing the conditions for one or more maxima to occur in the height distribution of electron concentration, for arbitrary height distributions of temperature and recombination coefficient. A numerical treatment is given for splitting into F_1 and F_2 layers. The occurrence of splitting with uniform temperature distribution would imply a highly improbable value for the recombination coefficient. On the other hand, the presence of a warm layer within the F layer would lead to results consistent with the observations.

551.510.535 426
Absorption of Short Waves in the Ionosphere—É. Argence and K. Rawer. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 773–775; September 27, 1954.] Expressions are derived analytically for the value of the decrement δ_2 for values of the frequency above and below critical, assuming that collisions of electrons with ions are much more numerous than with neutral molecules.

551.510.535:523.16 427
High Latitude Ionospheric Observations using Extraterrestrial Radio Waves—C. G. Little. (PROC. I.R.E., vol. 42, pp. 1700–1701; November, 1954.) A preliminary account is given of observations made at College, Alaska, over a two-month period in the spring of 1954, of the effect of the ionosphere on the reception of cosmic radiation at 65 mc. Strong absorption of the diffuse rf background is most common at midday and least common during late evening; it is correlated with no-echo conditions and with variations of the geomagnetic field. Observations on localized radio sources indicate that scintillations are commoner in Alaska than in England. Scintillations were observed whenever marked 65-mc absorption occurred during transit of the localized source through the antenna beam; this indicates that the lateral distribution of ionization during periods of absorption is nonuniform. Study of the scintillations should provide information about irregularities in the ionosphere.

551.510.535:538.566 428
The Propagation of Long Electric Waves round the Earth in relation to Transmission Interference and Lightning Atmospherics—Schumann. (*See* 528.)

551.594.21 429
An Analysis of Electric Field after Lightning Discharges—Y. Tamura. (*Jour. Geomag. Geoelect.*, vol. 6, pp. 34–46; March, 1954.) The thundercloud is represented by a vertical dipole, and the emf in it is represented by a constant current generator. The electric field comprises components due to (a) the charge on each pole, which depends on the conductivity of the air, (b) space charge related to the charge on the poles, and (c) free decaying space charge released after a lightning discharge. Dissimilarity between the field recovery curves for near discharges and distant discharges are explained by this theory.

551.594.6:621.396.11.029.45 430
Theoretical Calculations of Field of Low-Frequency Electromagnetic Waves above Earth's Surface—Al'pert. (*See* 535.)

LOCATION AND AIDS TO NAVIGATION

621.396.93:551.594.6 431
The Accuracy of the Location of Sources of Atmospherics by Radio Direction-Finding—F. Horner. (*Proc. IEE*, part III, vol. 101, pp. 383–390; November, 1954.) An investigation of the accuracy of the United Kingdom network of twin-channel cathode-ray direction finders operating at frequencies near 10 kc is reported. Hilly terrain and buried cables have caused errors of several degrees at some stations. Instrumental errors are small, apart from polarization errors, which may be as high as 2 degrees or 3 degrees in summer daytime and rather higher in winter. An estimate has been made of the accuracy of storm location; the potential accuracy depends not only on bearing errors, station spacing, and storm distance, but also on the number and interpretation of the observations. With the existing network and technique, the probable error in position of a storm center distant 1,000 km is about 20 km on a summer day, 50 km on a winter day, and 100 km by night.

621.396.933 432
Navaglobe-Navarho Long-Range Radio Navigational System—C. T. Clark, R. I. Colin, M. Dishal, I. Gordy and M. Rogoff. (*Elec. Commun.*, vol. 31, pp. 155–166; September, 1954, *Convention Record I.R.E.*, vol. 2, part 5, pp. 88–97; 1954.) The navaglobe system developed for the U.S.A.F. is described; reliability is achieved by operating at a frequency of about 100 kc, and tolerable signal /noise ratio at long range by using a bandwidth of 20–100 cps. The service is omnidirectional and the airborne bearing indications are automatic. Arrangements for reducing the effects of atmospheric noise are described. Performance on transcontinental and transatlantic tests is reported. Plans to complete the system by incorporating phase-comparison distance-measurement equipment are discussed.

621.396.933 433
Fixed-Beam Aircraft Approach System—R. A. Hampshire. (*Elec. Commun.*, vol. 31, pp. 189–197; September, 1954, *Convention Record I.R.E.*, vol. 2, part 5, pp. 77–83; 1954.) A system designed for the U.S.A.F. and compatible with existing instrument-approach installations consists of a localizer operating at a frequency between 108 and 112 mc and glide-slope equipment operating at a frequency between 329 and 335 mc, both using modulation at 90 and 150 cps. The localizer is of dual-beam type

621.396.96 434
Influence of the Polarization of the Radiated Waves on Radar Detection—G. Pircher. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 756–757; September 27, 1954.] Analysis for elliptically polarized radiation is based on a polarization factor which is given by the scalar product of unit vectors representing respectively the transmitter and receiver polarizations. It is shown that for one particular polarization of the radiated waves the echo is annulled, and that for any obstacle there is one particular polarization which gives maximum echo power.

621.396.962.23 435
Radar offers Solution to Midair Plane Collisions—J. Q. Brantley. (*Electronics*, vol. 27, pp. 146–150; November, 1954.) Early indication of collision danger is obtained by use of cw Doppler technique to measure closure rate, deviation of closure rate and deviation of closure angle. Suitable circuits are described.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 436
Ultimate Vacuum in a Vacuum-Enclosed Ionization Gauge—L. J. Varnerin, Jr., and D. White. (*Jour. Appl. Phys.*, vol. 25, pp. 1207–1208; September, 1954.) Brief report of measurements made with a Bayard-Alpert gauge.

533.583:[546.821+546.831 437
Investigation of Absorption Capacity of Nondispersal [contact] Getters—E. Lyubovskaya and A. Ravdel'. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1392–1400; August, 1954.) The sorption of hydrogen and oxygen by Zr getters and by Ti getters containing some Th was investigated in the range between room temperature and $\sim$700 degrees C. Results, which are presented graphically, indicate that both getters are suitable for use in tubes in the range investigated. The sorption capacity of Zr is highest at $\sim$500 degrees C. for H_2 and increases monotonically for O_2 above $\sim$400 degrees C.; for TiTh getters the corresponding temperatures are 400 degrees–600 degrees C. and $>$500 degrees C., respectively. At optimum temperatures Zr absorbs more O_2 than does the TiTh getter; the reverse is true for the H_2 sorption. The rate of sorption was also investigated.

535.215 438
Photoconductivity—A Rose. (*Onde élect.*, vol. 34, pp. 645–651; August/September, 1954.) Discussion of photoconduction mechanism in normally poor conductors such as CdS single crystals, pointing out discrepancies between experimental results and simple theory, and illustrating how energy-level distributions in the forbidden bands can be determined by development of suitable theoretical models. See also 1005 of 1952.

535.215.1:546.482.31 439
Photoconductivity in Cadmium selenide—H. J. Dirksen and O. W. Memelink. (*Appl. Sci. Res.*, vol. B4, pp. 205–215; 1954.) Published theories are discussed and a new theory is proposed. Values of the material constants calculated from the theory are in good agreement with results of measurements on CdSe in its basic form and in the form of thin films.

535.37:546.561–31 440
Nature of Luminescence of Copper Oxide—Yu. I. Karkhanin and V. E. Lashkarev. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 97, pp. 1007–1010; August 26, 1954. In Russian.] A brief report is presented summarizing recent experimental results. It is concluded that Cu vacancies do not luminesce by themselves, but that luminescence is produced at those centers which force the exciton to radiate on annihilation. The wavelength of the infrared luminescence of Cu_2O is about 0.96 μ at 20 degrees C. and 0.93 μ at -180 degrees C.

535.376 441
Experiments on Electroluminescence—J. F. Waymouth and F. Bitter. (*Phys. Rev.*, vol. 95, pp. 941–949; August 15, 1954.) Observations on green-luminescent ZnS:Cu:Pb and yellow-luminescent ZnS:Cu:PB:Mn phosphors are reported. The optical and electrical response of the materials to applied electric fields was studied for dispersions in lucite and for individual particles. Luminescence in alternating fields is restricted to spots very much smaller than the individual particles, and occurs only once per cycle. The results indicate that the most important excitation process involves removal of electrons from lumines-

cent centers, luminescence occurring when the state of the field allows the electrons to return and recombine.

537.227:546.431.811.824–31 442

Ferroelectric Properties of Solid Solutions of Barium Stannate in Barium Titanate—G. A. Smolenski and V. A. Isupov. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1375–1386; August, 1954.) Solid solutions containing up to 30 per cent molar of $BaSnO_3$ were investigated. Measurements were made of the dielectric constant in weak fields, the resonance and antiresonance frequencies and the fractional elongation $\Delta l/l$ over the temperature range -150 degrees to $+150$ degrees C. In several cases, the dependence of the dielectric constant on the field strength (up to $\sim$8 kv/cm) was also determined. Phase transitions were also investigated. The preparation of the specimens and experimental details are described and results are presented graphically and in some cases also in tabulated form.

537.227:546.431.817.824–31 443

X-Ray Study of Structure of (Ba, Pb) TiO_3 Solid Solutions—E. G. Fesenko and A. G. Slabchenko. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1288–1290; July, 1954.)

537.227:546.431.824–31 444

Pressed Barium Titanate Piezoelectric Materials—A. I. Kogan and M. M. Kitaigorodski. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1371–1374; August, 1954.) The dielectric constant and the piezoelectric modulus d_{11} were determined experimentally for polarized (Ba, Pb)TiO_3 specimens containing up to 18 per cent Pb. The dielectric constant lies between 40 and 95 and is practically constant at frequencies between 50 kc and 12 mc; d_{11} varies between 0.6×10^{-7} and 10×10^{-7} cgse. The effect of temperature and length of the thermal pre-treatment was investigated. Results are tabulated.

537.227:546.431.824–31 445

Domain Formation and Domain Wall Motions in Ferroelectric $BaTiO_3$ Single Crystals—W. J. Merz. (*Phys. Rev.*, vol. 95, pp. 690–698; August 1, 1954.) Continuation of work reported previously (759 of 1953). Nucleation and growth of the domains were studied as a function of applied electric field and temperature. The manner of growth of new domains on reversing the polarity of the applied field is discussed in relation to processes observed in ferromagnetic materials and to switching processes when the crystals are used as storage units. Compared with Fe, the walls in $BaTiO_3$ are much thinner and the energy density in the walls is somewhat greater. Since there is practically no sideways motion of the side walls, there is no crosstalk between different sets of electrodes on the same crystal even when they are spaced as closely as 0.01 cm.

537.227:546.431.824–31 446

Dielectric Constant and Loss Measurements on Barium Titanate Single Crystals while Traversing the Hysteresis Loop—M. E. Drougard, H. L. Funk and D. R. Young. (*Jour. Appl. Phys.*, vol. 25, pp. 1166–1169; September, 1954.) Permittivity measurements were made with small-amplitude sinusoidal voltages of frequency 100–600 kc superposed on a bias voltage switched at 10–200 cps. Results indicate that the frequency dependence of the permittivity and loss is consistent with a relaxation mechanism with a relaxation time of 5.5 μs. See also 3245 of 1954 (Drougard and Young).

537.227:621.375.5 447

Ferroelectrics and the Dielectric Amplifier—W. P. Mason and R. F. Wick. (Proc. I.R.E., vol. 42, pp. 1606–1620; November, 1954.) Properties of $BaTiO_3$ single crystals and ceramics are reviewed in relation to their use for capacitors, transducers, dielectric amplifiers, etc. By adding $PbTiO_3$ and $CaTiO_3$ the ceramics can be stabilized so as to be satisfactory for electromechanical filters and delay lines. For use in dielectric amplifiers and signal-storage units, single crystals are more suitable than ceramics by virtue of their narrower hysteresis loop. Experiments are briefly reported on dielectric amplifiers with crystals of thickness 0.002 inch operated below the Curie temperature; these crystals are saturated by a voltage of 10v, and their hysteresis loops do not vary with frequency up to 200 kc or higher. The amplifier has a gain of about 12 db for modulation frequencies up to 7 kc.

537.311.33:535.215 448

Photovoltaic Effect in InAs—R. M. Talley and D. P. Enright. (*Phys. Rev.*, vol. 95, pp. 1092–1094; August 15, 1954.) Report of measurements of the spectral variation of the photovoltaic effect in InAs *p-n* junctions.

537.311.33:536.21.022 449

Some Regularities in the Magnitudes of Thermal Conductivities of Semiconductors—A. V. Ioffe and A. F. Ioffe. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 97, pp. 821–822; August 11, 1954, In Russian.] The decrease in thermal conductivity with increase in atomic or molecular weight, and, in the case of ionic-lattice compounds, also with increase in the ratio of the atomic weight of the positive to that of the negative constituent element, is shown graphically. The semiconductors considered include alkali halides, AgCl, AgBr, TlCl and TlBr, diamond, Si, Ge, GaSb, and InSb.

537.311.33:537.312.8 450

Transverse Galvanomagnetic Effects in Semiconductors—J. Appel. (*Z. Naturf.*, vol. 9a, pp. 167–174; February, 1954.) The investigation reported by Johnson and Whitesell (2329 of 1953) is extended to magnetic fields of any strength. The value found for the magnetoresistance with strong fields and high carrier mobilities is much greater than that given by Harding's theory (*Proc. Roy. Soc. A.*, vol. 140, p. 205; 1933), and is in agreement with experimental results reported by Estermann and Foner (661 of 1951) and by Pearson and Suhl (166 of 1952). The two-band model is considered; the influence of intrinsic conduction on the magnetoresistance of Ge at room temperature and the change of sign of the Hall constant in some circumstances are discussed.

537.311.33:539.234 451

Evaporated Multiple Layers with Semiconductor Properties—J. C. M. Brentano and J. D. Richards. (*Phys. Rev.*, vol. 95, pp. 841–843; August 1, 1954.) Thin multiple-layer films were deposited on pyrex in vacuum by alternate evaporation of two metals not likely to form intermetallic compounds or alloys. The experimentally observed variation of resistivity with temperature is plotted for a film comprising 210 layers each of Fe and Pb; the temperature coefficient of resistivity is negative. For films of either Fe or Pb alone, produced under similar conditions, the temperature coefficients are negligible or positive. The significance of this difference is discussed in relation to the lattice structure.

537.311.33:546.28 452

Formation of Single-Crystal Silicon Fibers—E. R. Johnson and J. A. Amick. (*Jour. Appl. Phys.*, vol. 25, pp. 1204–1205; September, 1954.) Details are given of fibers formed when $SiCl_4$ diluted with a carrier gas is caused to react with Zn vapor at a temperature of 800 degrees–1,000 degrees C. The average diameter of the fibers is about 1 μ and the lengths are up to 1 cm.

537.311.33:546.28 453

Hyperfine Splitting in Spin Resonance of Group V Donors in Silicon—R. C. Fletcher, W. A. Yager, G. L. Pearson and F. R. Merritt. (*Phys. Rev.*, vol. 95, pp. 844–845; August 1, 1954.) Continuation of investigations reported previously [3254 of 1954 (Fletcher et al.)]. Samples with various concentrations of P, Ar and Sb have been studied; with all three donors, the multiplicity of resonance lines is replaced by a single narrow line when the concentration exceeds a certain value.

537.311.33:546.28 454

Precipitation of Impurties at Dislocations in Heat-Treated Silicon—S. Mayburg. (*Phys. Rev.*, vol. 95, pp. 838–839; August 1, 1954.) Results of experiments indicate that impurities are precipitated at dislocations when Si single crystals are heated at temperatures below 900 C. The heating was performed in vacuum by passing current through the specimens, and the effects were studied by measurements of the room-temperature conductivity after quenching. The critical temperature of 900 degrees C. is that at which plastic deformation is first observed in Si; a similar critical point is observed in Ge at about 500 degrees C. [2430 of September (Fuller et al.)].

537.311.33:[546.289+546.28+546.24 455

Effect of Surface Levels on Electrical Properties of Fine-Grain Films of Ge, Si, and Te—Ya. E. Pokrovski. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1229–1243; July, 1954.) The Hall coefficient and resistivity of films prepared by condensation in vacuum were determined over the temperature range from about -180 degrees C. up to several hundred degrees C. The grain dimensions, which were in the region of $10^{-5}-10^{-6}$ cm, were controlled by using thin quartz base plates at various temperatures. Results, which are presented graphically, indicate that the surface levels of Ge are acceptor levels filled, at room temperature, with electrons at a density of about 3.5×10^{11} cm^{-2}; the activation energies of the surface levels are about 0.004, 0.02 and 0.055 ev. The properties of Si films can be explained if it is assumed that an empty band lies just above a filled surface band, both near the middle of the forbidden band. Te has acceptor surface levels lying above the filled band.

537.311.33:546.289 456

Ionization Energies of Groups III and V Elements in Germanium—T. H. Geballe and F. J. Morin. (*Phys. Rev.*, vol. 95, pp. 1085–1086; August 15, 1954.) Experiments indicate that differences between the ionization energies of Group-III acceptors and Group-V donors in Ge are similar to, but less pronounced than, those previously reported for Si [*Bull. Amer. Phys. Soc.*, vol. 29, p. 22; 1954. (Morin et al.)].

537.311.33:546.289 457

Theory of Dislocations in Germanium—W. T. Read, Jr. (*Phil. Mag.*, vol. 45, pp. 775–796; August, 1954.) Theory is discussed quantitatively on the basis of a model due to Shockley (*Phys. Rev.*, vol. 91, p. 228; 1953). In *n*-type Ge a dislocation gives rise to a row of acceptors; the space charge which develops as these accept electrons has a controlling influence in determining the occupation of the remaining centers. Formulas are derived for the temperature variation of the average electron concentration in slightly deformed specimens. Experimental methods for testing the theory are suggested.

537.311.33:546.289:541.135 458

Self-Powered Semiconductor Amplifier—C. G. B. Garrett and W. H. Brattain. (*Phys. Rev.*, vol. 95, pp. 1091–1092; August 15, 1954.) A slice of *n*-type Ge with an In-alloyed emitter on one surface has its other surface exposed to a suitable electrolyte. Another electrode having a half-cell potential higher than that of Ge is also placed in the electrolyte. Connections are made to the In, to a gold-bonded contact on the Ge, and to the other electrode. Such a unit constitutes an amplifier incorporating its own power supply, multiplication of holes taking place at the Ge/electrolyte surface. Measured values between 1.4 and 2.0 have been obtained for the current gain. The cut-off fre-

quency is comparable to that of a transistor of similar dimensions.

537.311.33:546.3-1-28-289 459
Energy Gap of Germanium-Silicon Alloys—A. Levitas, C. C. Wang and B. H. Alexander. (*Phys. Rev.*, vol. 95, p. 846; August 1, 1954.) Measured values of energy gap are shown graphically against alloy composition.

537.311.33:546.3-1-28-289 460
Speculations on the Energy-Band Structure of Ge-Si Alloys—F. Herman. (*Phys. Rev.*, vol. 95, pp. 847–848; August 1, 1954.)

537.311.33:546.46-31 461
Electrical Conductivity of Magnesium Oxide Single Crystals—E. Yamaka and K. Sawamoto. (*Phys. Rev.*, vol. 95, pp. 848–850; August 1, 1954.) Results of experiments indicate that the 2.3-ev energy level available to the charged carriers originates from excess Mg and not from excess 0 atoms.

537.311.33:546.682.18 462
Optical Properties of Indium Phosphide in the Infrared—F. Oswald. (*Z. Naturf.*, vol. 9a, p. 181; February, 1954.) Report of measurements over the wavelength range 0.8–15.2 μ on an n-type specimen of resistivity 0.1 Ω cm. A sharp absorption edge is found corresponding to an energy gap of 1.25 ev. Absorption observed at wavelengths >14.5 μ may be due either to an impurity or to the lattice vibrations.

537.311.33:[546.682.86+546.289 463
Analysis of Magnetoresistance and Hall Coefficient in p-Type Indium-Antimonide and p-Type Germanium—T. C. Harman, R. K. Willardson and A. C. Beer. (*Phys. Rev.*, vol. 95, pp. 699–702; August 1, 1954.) Existing theory is re-cast in a form convenient for the analysis of experimental results. For p-type InSb good agreement is obtained between experimental and theoretical results; in the case of p-type Ge the theory is not capable of explaining the experimentally observed variations of Hall coefficient with magnetic field or of magnetoresistance with temperature.

537.311.33:546.722-31 464
Application of R.F. Absorption in Iron Oxides to the Determination of the Activation Energy of their Lattice Defects—B. Hagène. (*Jour. Phys. Radium*, vol. 15, pp. 583–584; July/September, 1954.) Continuation of work reported previously [2328 of 1953 (Freymann et al.)]. Examination of experimental curves in the light of the formula relating frequency of maximum Debye absorption to absolute temperature indicates that for iron oxide mixtures in group the activation energy is $\sim$0.1 ev, probably related to defects due to the presence of Fe_3O_4. For another group the activation energy is 0.5–0.9 ev, possibly corresponding to an electronic phenomenon. The activation energy of αFe_2O_3 is 0.4 ev and of γFe_2O_3 0.07 ev.

537.311.33:548.53 465
Solute Redistribution by Recrystallization—R. G. Pohl. (*Jour. Appl. Phys.*, vol. 25, pp. 1170–1178; September, 1954.) Theoretical investigation of the modification of the distribution of impurities in a melt of metal or alloy during solidification. Results are presented graphically and tabulated, and are compared with previously published findings.

537.311.33:621.314.7 466
Transistor Electronics: Imperfections, Unipolar and Analog Transistors: Parts 2 and 3—W. Shockley. [*Proc. IRE* (*Australia*), vol. 15, pp. 194–201 and 228–234. August and September, 1954.] Reprint, see 746 of 1953. Part 1: 3600 of 1954.

537.582 467
Thermionic Constants of Metals and Semiconductors: Part 4—Monovalent Metals (continued)—S. C. Jain and K. S. Krishnan. (*Proc. Roy. Soc. A*, vol. 225, pp. 159–172; August 31, 1954.) Regarding the electrons in the metal and in the vapor phase as a single-component thermodynamic system, the temperature coefficient of the work function ϕ corresponds to an apparent lowering of Richardson's constant A by about 8–10 per cent. This agrees with experimental results. The effects of thermal expansion of the lattice and thermal agitation of the atoms on ϕ practically cancel each other, so that the observed variation of ϕ with temperature may be regarded as due to electronic specific heat alone. Part 3: 2700 of 1953.

538.221 468
The Spontaneous Magnetization of Alloys: Part 1—Copper Nickel Alloys—D. J. Oliver and W. Sucksmith. (*Proc. Roy. Soc. A*, vol. 219, pp. 1–17; August 11, 1953.) Report and discussion of measurements made to determine the variation of spontaneous magnetization with temperature. A magnetothermal method gives a curve below that obtained from purely magnetic measurements. The relation of the magnetocaloric temperature rise to the square of the magnetization is nonlinear above the Curie point.

538.221 469
Temperature of Heating in Thermomagnetic Treatment of Magnico Alloy—G. F. Golovin and A. A. Shekalov. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1503–1507; August, 1954.) The effect was investigated experimentally of various heat treatments of the alloy (50.5 per cent Fe—23.6 per cent Co—13.6 per cent Ni—8.6 per cent Al—3.0 per cent Cu—0.05 per cent C) on the structure and magnetic properties. Results are presented graphically and microphotographs of surface structure are shown.

538.221 470
A Theory of Domain Creation and Coercive Force in Polycrystalline Ferromagnetics—J. B. Goodenough. (*Phys. Rev.*, vol. 95, pp. 917–932; August 15, 1954.) "Granular inclusions, grain boundaries, lamellar precipitates, and the crystalline surface have been examined as possible nucleation centers for domains of reverse magnetization in ferromagnetic materials. It is concluded that the surface density of magnetic poles at the grain boundaries is the most common source of nucleation energy in polycrystalline materials. The concept of nucleation of domains of reverse magnetization has led to a calculation of three more terms which may contribute to the coercive force in polycrystalline materials, viz., a grain-boundary, a lamellar-precipitate, and a domain-wall-surface-tension contribution. The theoretical predictions are compared with several old experiments. New insight is gained on the problem of stress sensitivity of polycrystalline ferromagnetics and B-H loop shape."

538.221:538.114 471
Theory of Ferromagnetic Resonance and Results of Experimental Investigation—K. H. Reich. (*Z. angew Phys.*, vol. 6, pp. 326–338; July, 1954.) A survey of experimental and some theoretical work, with particular reference to Kittel's theory. 83 references include work up to October 15, 1953; several more recent references are given in the addendum.

538.221:538.652 472
Investigation of Magnetostriction of Iron-Nickel Alloy in Strong Magnetic Fields—G. P. D'yakov and R. A. Reznikova. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 97, pp. 633–634; August 1, 1954. In Russian.] An experimental investigation of wires of composition 41 per cent Fe and 59 per cent Ni is reported. Plotting $\lambda_s - \lambda$ against H^{-2}, where λ_s is the saturation magnetostriction and H the magnetizing field, results in a straight line the slope of which is 32 K^2 $\lambda_s/35$ J_s^2, where K is a magnetic anisotropy constant and J_2 the saturation magnetization. The result is briefly discussed.

538.221:538.652 473
Magnetostriction of Ferromagnetic Alloys based on Manganese—D. I. Volkov. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 97, pp. 809–811; August 11, 1954. In Russian.] The materials investigated comprised the binary Mn alloys with Sb, Bi and Sn, and alloys of the type Cu-Mn-X, where X is Al, Sn, or Bi. Several of these are highly magnetostrictive both at room temperature and at -195 degrees C. The magnetostriction of 31.2 per cent Mn—68.8 per cent Sb rises to a value more than twice that of Ni at saturation; the longitudinal magnetostriction of 30.9 per cent Cu—14 per cent Mn—55.1 per cent Bi is about -140×10^{-6} at saturation, a value nearly equal to that for the best Pt-Fe alloy investigated by Kussmann and Rittberg (2788 of 1950). Experimental curves are given for several alloys.

538.221:538.652 474
A Peculiarity of Magnetostrictive Properties of Mn-Sn Ferromagnetic Alloys—D. I. Volkov and V. I. Leont'ev. [*Compt Rend. Acad. Sci.* (*URSS*), vol. 97, pp. 995–997; August 21, 1954.] Results are reported of an experimental determination of the longitudinal, transverse and volume magnetostriction in Mn-Sn alloys containing between 42 and 48 per cent Mn, at temperatures of -78 degrees and -195 degrees C. All three magnetostriction coefficients were found to be negative at saturation.

538.221:621.318.134 475
The Spontaneous Magnetization of Alloys and Compounds: Part 2—Ferrites—C. A. Clark and W. Sucksmith. (*Proc. Roy. Soc., A*, vol. 225, pp. 147–159; August 31, 1954.) Magnetic and magnetocaloric measurements have been made on Mn ferrite and Mg-Zn ferrite. Values of spontaneous magnetization obtained by three extrapolation techniques are similar; this agreement is maintained at temperatures about the Curie point, in contrast to the results for ferromagnetic materials [part 1: 468 above (Oliver and Sucksmith)]. The saturation magnetization of both ferrites and the reduced-magnetization/reduced-temperature curve for Mn ferrite are in satisfactory agreement with Néel's theory.

538.221:621.318.134 476
Magnetic Ferrites: New Materials for Modern Applications—V. E. Legg and C. D. Owens. (*Elec. Eng.* (*N.Y.*), vol. 73, pp. 726–729; August, 1954.) Review of properties and applications of different ferrites, with a note on manufacturing processes. Typical characteristics of NiZn, MnZn and Ni ferrites are tabulated. Q values and applications in different frequency ranges are listed.

538.247 477
Investigation of Hysteresis of Ballistic Demagnetization Factor—A. P. Lyustrova and V. A. Lipatova. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1513–1519; August, 1954.) An experimental investigation is reported. Soft and hard ferromagnetic specimens in prismatic form were used.

621.315.61:537.533.8 478
A Method of investigating the Secondary Emissive Properties of Insulators—C. N. W. Litting. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 289–293; August, 1954.) The method is limited to surfaces with $\delta>1$. The insulator under investigation is in contact with a backing plate to which is applied a square wave whose mark /space ratio is varied to obtain zero net loss of charge from the surface over a complete cycle. The mark/space ratio of this waveform is then used to compute the value of δ. Results obtained from simultaneous measurements using this method and a dc one show very good agreement. Difficulties are discussed. The method may also be used to determine the energy distribution of the secondary electrons.

621.315.612:537.226.31 479
Dielectric Losses in High-Frequency Ceramics—N. P. Bogoroditski and I. D. Fridberg. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1194–1204; July, 1954.) A discussion of previously published experimental results is used as a basis for classifying dielectric loss mechanisms; seven main types are tabulated and the conditions for these losses to be low or high are stated.

621.315.612:621.385.032 480
Advances in Ceramics related to Electronic Tube Developments—L. Navias. (*Jour. Amer. Ceram. Soc.*, vol. 37, pp. 329–350; August 1, 1954.) A detailed study is made of the physical properties of alumina, forsterite, zircon-talc and steatite in relation to the requirements of tubes for operation in the 3-cm waveband. The development of sealing techniques is described, and details are given of methods of making ceramic/metal seals using (a) Mn-Mo, (b) MoO_3 or (c) TiH_2 as the metallizing agent.

621.315.613.1 481
Three Forms of Synthetic Micas—T. B. Merrill, Jr. (*Materials and Methods*, vol. 40, pp. 80–83; August, 1954.) Discussion of the preparation, properties and uses of material obtained by (a) glass-bonding finely powdered mica, (b) reconstituting small flakes into sheets by calendering, or (c) hot-pressing finely powdered mica into homogeneous blocks.

621.791.3:546.621 482
Soldering Aluminum—J. C. Bailey and J. A. Hirschfield. [*Research* (*London*), vol. 7, pp. 320–326; August, 1954.] A review is presented of methods, materials, techniques and applications. Tables are given of the composition of various solders and their melting points, and of results of breaking and tensile strength tests on soldered joints. Solder embrittlement and corrosion are also discussed.

MATHEMATICS

517 483
On the Behavior of the Solutions of the Differential Equation $\Delta U=F(x, U)$ in the Neighborhood of a Point—C. Müller. (*Commun. pure Appl. Math.*, vol. 7, pp. 505–515; August, 1954.) An equation involved in the theory of em oscillations is investigated.

517.9 484
An Extended Use of Perturbation Theory—R. N. Gould and A. Cunliffe. (*Phil. Mag.*, vol. 45, pp. 818–822; August, 1954.) A method of solving partial differential equations arising in boundary-layer problems is given.

517.7 485
Handbook of Elliptic Integrals for Engineers and Physicists [Book Review]—P. F. Byrd and M. Friedman. Publishers: Springer, Berlin, Germany, 1954, 355 pp., 36 DM. (*Arch. elekt. Übertragung*, vol. 8, p. 403; September, 1954.)

518.3 486
Leitfaden der Nomographie [Book Review]—W. Meyer zur Capellen. Publishers: Springer, Berlin, Germany, 1953, 178 pp., 17.40 DM. (*Z. angew. Phys.*, vol. 6, p. 384; August, 1954.) The theoretical bases of nomography are presented and numerous examples and applications are given.

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74)+529.786]:538.569.4 487
Atomic Clocks and Frequency Standards on an Ammonia Line: Parts 2 and 3.—K. Shimoda. [*Jour. Phys. Soc.* (*Japan*) vol. 9, pp. 558–575; July/August, 1954.] Use of the Zeeman and Faraday effects for eliminating causes of error in atomic clocks is examined. The nonreciprocal transmission characteristics of a waveguide absorption cell in an axial magnetic field are investigated theoretically in relation to the design of a highly accurate clock. An arrangement is described which makes use of Stark, source and Zeeman modulation and is free from errors due to reflections in the microwave line. Preliminary results using a cell of length 1 m indicate that accuracy within 10^{-9} is obtainable.

621.3.018.41(083.74):621.373.421.13 488
Audio and R.F. Secondary Frequency Standard—R. C. Moses. (*Radio-Electronic Eng.*, vol. 23, pp. 16–18, 30–31; August, 1954.) The frequency standard supplies sine-wave and square-wave outputs at five frequencies between 1 kc and 100 kc, and harmonics of a basic 100-kc oscillator up to 150 mc. The 100-kc master oscillator is maintained at constant temperature and operated at very low power level. Supplementary crystal-stabilized 1-mc and 10-mc oscillators are used. The frequency divider comprises cascaded decade-scaling circuits with stability dependent primarily on tuned *LC* circuits.

621.316.84(083.74) 489
Enclosed Standard Resistors—A. Schulze and H. Zacherl. (*Z. angew. Phys.*, vol. 6, pp. 324–326; July, 1954.) The construction of the Au-Cr resistor [772 of 1953 (Schulze and Ericke)] and of the manganin resistor is modified to make it watertight. Helium is found to provide the best atmosphere for the Au-Cr type, argon or nitrogen for the manganin type.

621.317.3 490
Measurement Methods—(*Fernmeldetech. Z.* vol. 7, pp. 379–436; August, 1954.) A special number including the following papers:

"Measurements on Coaxial Cables,"—A. Knacke (pp. 387–393.) The pulse reflection method is discussed.

"High-Accuracy Direct-Indicating Frequency Meter,"—J. Hacks (pp. 394–398). Equipment using counter circuits is described.

"Development Trends in Impedance Meters for Very High Frequencies,"—R. Eichaker (p. 399).

"Ratio Meter for Determination of Admittance particularly in the U.S.W. Range,"—H. Fricke (p. 400).

"Direct Indication of Envelope Delay in the Frequency Range 100 kc/s–5 Mc/s,"—W. Kaiser and H. Wilde (pp. 401–405). Variations of envelope transmission time down to 10^{-9} seconds are determined from very accurate phase measurements.

"Investigations with a Direct-Indicating Envelope-Delay Meter in the Range 60–90 Mc/s,"—H. Reiner (pp. 406–409). Oscillograms are shown of phase-delay characteristics of various quadripoles, if amplifiers and imperfectly terminated cables.

"Methods of Measurement of Impedance at Frequencies up to 10 kMc/s,"—A. Jaumann (pp. 410–418). Methods using bridges and directional couplers for measurement of impedance or reflection coefficient are described.

"Noncontacting Plungers for Microwave Instruments,"—J. Deutsch and O. Zinke (pp. 419–424). New designs give increased bandwidth and reduced losses.

"Time-Expanding Methods for Cathode-Ray Oscillographs,"—J. Czech (pp. 425–430).

621.317.3:621.314.7 491
Stability Considerations in the Parameter Measurements of V.H.F. Point-Contact Transistors—D. E. Thomas. (Proc. I.R.E., vol. 42, pp. 1636–1644; November, 1954.) In early forms of measurement equipment for transistors the α cut-off frequencies involved were low enough for lead impedances and parasitic shunt capacitances to be neglected. Modified forms of equipment are discussed suitable for dealing with later types of transistor having α cut-off frequencies in the vhf range. Negative-resistance collector regions may give rise to discontinuities in the static characteristics. Simple circuits are shown for characteristic-curve tracers with either oscilloscope or X-Y recorder presentation.

621.317.3:621.396.674.1 492
Generation of Standard Fields in Shielded Enclosures—F. Haber. (Proc. I.R.E., vol. 42, pp. 1693–1698; November, 1954.) With the known method of calibrating loop antennas by arranging the loop in the vicinity of a radiating wire inside an enclosure, the boundary conditions in the formulas involved are satisfied for an infinite set of images. Errors are introduced if consideration is limited to one or two images; errors are also caused by frequency limitations. New formulas are derived which give the field strength accurate to within 1 per cent for low and medium frequencies. A method of estimating the error for high frequencies is discussed.

621.317.331 493
An Absolute Measurement of Resistance by Albert Campbell's Bridge Method—G. H. Rayner. (*Proc. IEE*, part IV, vol. 101, pp. 250–257; August, 1954. Digest, *ibid.*, part II, vol. 101, pp. 574–575; October, 1954.) Two determinations of the unit of resistance in terms of mutual inductance and frequency were made during 1951 using Campbell's bridge method. They gave the same value within 1 part per million, and the mean value agreed with that found in 1936 to the same accuracy. The value obtained for the ohm as maintained by the group of coils forming the reference standard at the National Physical Laboratory is: 1 N.P.L. ohm = 1.000 002 ±0.000 015 absolute ohm.

621.317.331:621.385.032.216 494
Sensitive Apparatus for the Measurement of Cathode Parasitic Resistance—Sevin. (*See* 598.)

621.317.333.6.029.62 495
A Breakdown Cell for Measuring the Dielectric Strength of Solids at 100 Megacycles—L. J. Frisco and J. J. Chapman. (*Rev. Sci. Instr.*, vol. 25, pp. 733–737; August, 1954.) The cell is liquid-filled and contains the specimen under test, a high-voltage resonant circuit, and a built-in probe for measuring voltage. The calibration of the voltage-measurement network is discussed.

621.317.335:621.317.336.029.6 496
The Computation of the Complex Dielectric Constant from Microwave Impedance Measurements—J. P. Poley. (*Appl. Sci. Res.*, vol. B4, pp. 173–176; 1954.) Diagrams supplementing that developed by Benoit (935 of 1950) are presented for facilitating determination of the complex dielectric constant.

621.317.335:621.372.413 497
Dielectric Measurements with H_{01n} Resonant Cavities having Appreciable Loading—J. L. Farrands. (*Proc. IEE*, part III, vol. 101, pp. 404–406; November, 1954.) "Modifications to the usual formulas for dielectric measurements in H-mode cavities are suggested to allow for damping due to coupling irises. The correction permits the use of low-level sources."

621.317.335.2.029.4 498
The Measurement of Capacitance at Low Frequencies—P. Phillips. [Proc. IRE (Australia), vol. 15, pp. 191–193; August, 1954.] "The fundamental formulas relating the series and parallel equivalent circuits of a lossy capacitor are given and some methods of measuring capacitor properties at low frequencies are discussed."

621.317.335.3(083.74) 499
Capacitor with Definite Loss Angles for Checking Bridge Measurements at Power Frequency—A. W. Stannett. (*Jour. Sci. Instr.*, vol. 31, p. 304; August, 1954.) Description of improvements to the synthetic "lossy" capacitor constructed to represent a filled oil cell and

required in connection with the international comparison of power-factor measurements.

621.317.353+621.396.822]:621.316.8 500
New Resistor Voltage-Coefficient Tester—Rosenthal and Louis. (*See* 355.)

621.317.42:621.385.832 501
The Use of an Electron Beam for the Accurate Measurement of Alternating Magnetic Field Strengths—S. E. Barden and K. Phillips. (*Proc. IEE*, part II, vol. 101, pp. 441–449; August, 1954. Digest, *ibid.*, part III vol. 101, p. 406; November, 1954.) A method of making measurements relative to a standard field uses a lv electron-optical system which produces a beam if the magnetic field in which it is placed is less than or of the order of 0.01G. The measurement is based on the voltage pulse produced at field zero. The operation of the instrument in time-varying fields such as those in particle accelerators is discussed.

621.317.44 502
An Automatic Plotter for Magnetic Hysteresis Loops—H. McG. Ross. (*Proc. IEE*, part II, vol. 101, pp. 417–427; August, 1954. Discussion, pp. 427–430.) An instrument is described with which measurements are made under substantially zero-frequency conditions, the period of the cycle being nearly one minute. The magnetic flux is detected by a search coil on the sample, the emf being amplified and integrated by a variable-speed motor on the plotting table, whereby the displacement of the pen is a measure of the magnetic induction. Calibration is in terms of voltage and time.

621.317.7:621.373.421.1 503
Small-Amplitude Oscillator for Capacitive Measurement of the Deflection of Sensitive Rotating Systems—J. W. Hiby and K. G. Müller. (*Z. angew. Phys.*, vol. 6, pp. 361–363; August, 1954.) In a modified form of the capacitance micrometer using a quartz resonator, described e.g. by Dye (*Proc. Phys. Soc.*, vol. 38, pp. 399–457; 1926.), the oscillation amplitude is made very small so that the mechanical force on the oscillator plate is practically zero. Variation from the nominal frequency of 3.17 mc is not more than 5 parts in 10^6. Spontaneous amplitude fluctuations occur when the amplitude falls below 15 mv_{eff}.

621.317.723 504
A General-Purpose Electrometer—R. M. Fry. (*Jour. Sci. Instr.*, vol. 31, pp. 269–271; August, 1954.) An instrument based on the Townsend null method, but incorporating automatic adjustment of the compensating voltage, is described. It is suitable for free-air-chamber work and for extrapolation-chamber work over the current range 10^{-12}–10^{-9} a, and can be adapted for use in a bridge circuit for capacitance measurement.

621.317.729 505
Some Applications of the Electrolytic Tank to Engineering Design Problems—H. Diggle and E. R. Hartill. (*Proc. IEE*, part II, vol. 101, pp. 349–364; August, 1954. Discussion, pp. 364–368.) The solution of problems relating mainly to electric and magnetic fields is discussed.

621.317.733:621.317.4 506
A New Method for Measurement of Complex Susceptibility by Magnetic Resonance Absorption in the U.S.W. Range—W. Müller-Warmuth. (*Naturwiss.*, vol. 41, p. 368; August, 1954.) A bridge method is used in which the balance is highly sensitive to the absorption. A special heterodyne receiver is used as a null indicator instrument.

621.317.743:621.372.56 507
A Radio-Frequency Transformer Attenuator for use with a Level Recorder—C. G. Mayo and R. E. Jones. (*Proc. IEE*, part III, vol. 101, pp. 401–403; November, 1954.) An attenuator for operation at 10.7 mc is described which can be substituted for the potentiometer of a commercial high-speed level recorder without requiring modification of the latter. A method of recording field strength is indicated using the attenuator and level recorder with a receiver having an IF of 10.7 mc.

621.317.784.029.63 508
U.H.F. Meter measures Low Power Levels—R. L. Bailey and J. B. Quirk. (*Electronics*, vol. 27, pp. 159–161; November, 1954.) A coaxial-line device is described for measuring power in the range 10–100 μw at frequencies up to 900 mc with an insertion loss <0.8 db.

621.37/.39.001.4 509
Pan-climatic Testing—G. W. A. Dummer, S. C. Schuler and J. E. Green. (*Wireless World*, vol. 60, pp. 598–600; December, 1954.) Information compiled for the Ministry of Supply K. 114 series of test schedules for Service equipment is summarized. The tests cover combinations of extreme climatic conditions and mechanical stresses. A list is given of common faults in electronic equipment as indicated by this series of tests.

621.373.4.029.64 510
New 7-11-kMc/s Signal Generator yields Valuable Design Hints—A. Fong. (*Tele-Tech*, vol. 13, pp. 92–93, 184; August, 1954.) An account is given of some of the problems encountered in the design of a signal generator suitable for pulse work, slotted-line work and applications requiring a sawtooth FM signal. The oscillator uses a Type-5721 reflex klystron with a $\frac{3}{4}\lambda$ coaxial-line resonator. The frequency range is covered by means of two repeller modes: the $3\frac{3}{4}$ mode for the 7–9 kmc range, the $4\frac{3}{4}$ mode above 9 kmc. A special plunger is used to suppress the $\lambda/4$ resonator mode. A waveguide operated beyond cut-off is used as output attenuator. The power-output monitor is also briefly discussed.

621.397.61.001.4 511
Color Test Technique for TV Broadcasters—J. W. Wentworth. (*Electronics*, vol. 27, pp. 120–123; November, 1954.) Methods of testing the amplitude/frequency, phase/frequency, differential-gain and differential-phase characteristics of transmitters are outlined. Color-bar and dot-pattern generators are described.

621.397.62.001.4 512
Measurements on Television Receivers: Part 5—Measurements on the Deflection System—O. Macek. (*Arch. tech Messen*, no. 224, pp. 205–208; September, 1954.) Part 4: 3634 of 1954.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

526.2:535.22.082.5/.7 513
The Geodimeter—an Electronic "Eye" for Measuring Distance—C. E. Granqvist. (*Tele-Tech*, vol. 13, pp. 68–69, 148; September, 1954.) Bergstrand's Kerr-cell method of measuring the velocity of light (3396 of 1952) is applied to the converse problem of distance measurement.

535.24:778 514
A Simple Integrating Light Meter—W. Hartnagel. (*Z. angew. Phys.*, vol. 6, pp. 310–313; July, 1954.) The measurement is affected by means of a photocell, the current from which periodically discharges a capacitor; the latter is automatically recharged by a relaxation oscillator which simultaneously actuates a counter.

621.316.718:534.85:621.94 515
Recorder-Controlled Automatic Machine Tools—E. W. Leaver and G. R. Mounce. (*Electronics*, vol. 27, pp. 124–128; November, 1954.) Description of the operation of an engine lathe in which the carriage feed and cross feed are actuated by motors controlled by signals magnetically recorded on wire or tape. A sinusoidal reference oscillation is used.

621.317.39:61 516
Electrical Techniques in Medicine—C. N. Smyth. (*Trans. Soc. Instr. Tech.*, vol. 6, pp. 84–95; June, 1954. Discussion, pp. 95–96.) A survey is presented of electrical instruments with particular reference to the measurement of pressure and strain. About 100 applications are classified in nine main groups and the principles of operation of several instruments, such as the ultrasonic microscope, are briefly described. 63 references.

621.383.2:778.37 517
Application of the Electron-Optical Converter for Photography of High-Speed Processes—M. P. Vanyukov and E. V. Nilov. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1209–1218; July, 1954.) Application of the A.E.G. electrostatically focused three-electrode image-converter to the photography of spark discharges of duration 0.4–2 μs is discussed. Image distortion observed with photocathode illumination $>10^6$ lux is related to the field distribution in front of the photocathode and the build-up of the space charge.

621.384.611 518
Alternating Gradient Focusing of Cyclotron External Beam—E. L. Hubbard and E. L. Kelly. (*Rev. Sci. Instr.*, vol. 25, pp. 737–739; August, 1954.) Description of an arrangement comprising three alternately converging and diverging electrostatic lens elements of hyperbolic cross section for focusing the $\frac{3}{8}$-MeV external proton beam for the 20-inch Berkeley cyclotron.

621.385.833 519
Requirements contributing to the Design of Devices used in correcting Electron Lenses—G. D. Archard. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 294–299; August, 1954.) Voltage requirements and positioning tolerances are discussed.

621.385.833 520
Experimental Investigation of Cylindrical Magnetic Electron Lenses—V. M. Kel'man, D. L. Kaminski and S. Ya. Yavor. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1410–1427; August, 1954.) Formulas derived, assuming the lens field to be practically equal to that produced by equal currents flowing in opposite directions along two parallel infinitely long wires, were tested experimentally. The calculated and measured values were in good agreement. Results for one lens and for a system of two cylindrical lenses are presented graphically and the apparatus used is briefly described.

621.385.833 521
The Theoretical Resolving Power of an E.S. Immersion Objective—A. Septier. (*Jour. Phys. Radium*, vol. 15, pp. 573–581; July/September, 1954.) Calculations based on geometrical optics give a limiting value of the order of 30 mμ for the resolving power. This might be reduced to about 10 mμ by special grid design, but spherical aberration is then of the same order of magnitude. Calculations based on the distribution of electron density in planes adjacent to that of the Gaussian image give a theoretical limiting value of the order of 1 mμ.

621.385.833:537.533 522
Field-Emission Current Densities and Surface Field Strengths in Field-Emission Microscopes, and Methods for the Determination of the Radius and Shape of the Point, the Magnification and the Resolving Power—M. Drechsler and E. Henkel. (*Z. angew. Phys.*, vol. 6, pp. 341–346; August, 1954.)

621.387.424 523
The Efficiency of Halogen-Quenched Geiger-Counters for X Rays—D. van Zoonen.

(*Appl. Sci. Res.*, vol. B4, no. 3, pp. 196–204; 1954.)

621.387.424 524

Small-Diameter Geiger-Müller Counters with External Cathode—D. Blanc. (*Jour. Phys. Radium*, vol. 15, pp. 590–591; July/September, 1954.) Counters 6 mm in diameter, with methylal/argon filling, and having a Geiger threshold of 820 v are described.

621.398:621.396.65 525

Radiation Monitor for Atomic Explosions—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 38, pp. 116–118; August, 1954.) Data on the radiation intensity and other variables in the neighborhood of the explosion are automatically measured and transmitted by radio to a control station, where the operator is able to select one of several data stations and is also able to select the particular property on which to be informed. The vhf FM radio link and station interrogation system are based on standard equipment including a 0.25-w portable transmitter-receiver. Repeater stations capable of handling information from ten data stations are located at positions of high elevation.

681.81:621.3 526

Electronic Musical Instruments—W. Kwasnik. (*Elektrotech. Z., Edn.B*, vol. 6, pp. 305–312; August 21, 1954.) A historical survey from 1761 up to date includes brief descriptions of instruments, their principles of operation and circuits used. 31 references.

PROPAGATION OF WAVES

538.566:537.562:621.37 527

Analogues of an Ionized Medium—R. N. Bracewell. (*Wireless Eng.*, vol. 31, pp. 320–326; December, 1954.) The propagation of plane waves in ionized media can be represented by equations of the same form as those representing wave propagation in transmission lines. Equivalent circuits are presented for propagation in (a) free space, (b) an ionized medium, (c) an ionized medium, taking account of collisions, (d) the ionosphere, taking account of inhomogeneities. The analogy provides an explanation of the ability of a highly reflecting layer to transmit radiation if the medium beyond the reflecting layer satisfies certain conditions of resonance. Mechanical and waveguide analogs of the ionized medium are also mentioned.

538.566:551.510.535 528

The Propagation of Long Electric Waves round the Earth in relation to Transmission Interference and Lightning Atmospherics—W. O. Schumann. (*Z. angew. Phys.*, vol. 6, pp. 346–352; August, 1954.) Literature on long-wave propagation is surveyed. In order to explain the "slow-tail" phenomena of atmospherics, very low values of ionosphere conductivity must be assumed. On the other hand, in order to explain numerous observations of signals comprising multiply reflected pulses high conductivity must be assumed, and in some cases different reflection heights. Further observations and analysis are required.

621.396.11+551.510.53 529

Rocket Exploration of the Upper Atmosphere—Boyd, Seaton and Massey. (*See* 424.)

621.396.11 530

Velocity of Radio Waves—R. L. Smith-Rose. (*Wireless World*, vol. 60, pp. 590–591; December, 1954.) At the 11th General Assembly of the International Scientific Radio Union in 1954 the value of 299792±2 km was adopted for the velocity of em waves in vacuum. The circumstances leading to this decision are outlined.

621.396.11 531

The Extension of Sommerfeld's Formula for the Propagation of Radio Waves over a Flat Earth, to Different Conductivities of the Soil—H. Bremmer. (*Physica*, vol. 20, pp. 441–460; August, 1954.) Propagation over a mixed terrain is discussed, starting from an integral equation of the type used by Hufford (2311 of 1952). Such an integral equation can take account of different distributions of the dielectric constant and conductivity in the region between the point source and the receiving point. The solution for the special case of two adjacent homogeneous media is identical with Clemmow's solution (3688 of 1953). Expressions derived are suitable for numerical computations. The case of three adjacent media is also investigated.

621.396.11 532

Ray-Path Characteristics in the Ionosphere—G. Millington. (*Proc. IEE*, part IV, vol. 101, pp. 235–249; August, 1954. Digest, *ibid.*, part III, vol. 101, pp. 193–196; May, 1954.) A quicker and more accurate method of making magneto-ionic calculations is presented than that previously given (778 of 1952), based on the introduction of a new parameter. Propagation in the magnetic meridian plane is studied in detail, including the phenomenon described by Poeverlein of a cusp at the ray-path apex within a certain critical angle of incidence. The theory indicates that the effect of lateral deviation on direction finding is too small to be observable except under special conditions unlikely to occur in practice.

621.396.11:531.510.535 533

Some Notes on the Absorption of Radio Waves reflected from the Ionosphere at Oblique Incidence—W. J. G. Beynon. (*Proc. IEE*, part III, vol. 101, pp. 367–370; November, 1954.) Discussion on 1550 of 1954.

621.396.11:551.510.535 534

The Reflection and Absorption of Radio Waves in the Ionosphere—W. R. Piggott. (*Proc. IEE*, part III, vol. 101, pp. 367–370; November, 1954.) Discussion on 1679 of 1953.

621.396.11.029.45:551.594.6 535

Theoretical Calculations of Field of Low-Frequency Electromagnetic Waves above Earth's Surface—Ya. L. Al'pert. [*Compt. Rend. Acad. Sci.* (URSS), vol. 97, pp. 629–632; August, 1954. In Russian.] The propagation of lightning atmospherics at frequencies between about 0.5 and 15 kc is considered. Numerical results of field-strength calculations are presented graphically using a formula relating the field strength (in mv/m) at distances greater than about 50–100 km to the frequency, the antenna current (in a), the effective antenna and ionosphere heights and the distance (all in km), and functions of the reflection coefficient and complex permittivity of the ionosphere. Infinite conductivity of the earth's surface is assumed. The variation of the amplitude and of the phase velocity with distance is also shown graphically. Results are compared with experimental results of Chapman and Matthews (419 of 1954) showing the relation between relative amplitude and frequency at distances of 500, 1000 and 2000 km, and those of Weekes (1491 of 1950) showing the relation between signal strength × distance and distance; the frequency used in the calculation was near 15 kc, that in the latter experiments 16 kc. Good agreement was obtained with both sets of observations, indicating that the irregular properties of the field are principally due to complex interference effects rather than to irregularities of the transmission medium.

621.396.11.029.55:551.510.535 536

Ionospheric Absorption at Vertical and Oblique Incidence—G. McK. Allcock. (*Proc. IEE*, part III, vol. 101, pp. 360–367; November, 1954. Discussion, pp. 367–370.) Research on ionospheric absorption is briefly reviewed and an account is given of an experimental investigation made in New Zealand during 1949–1950. The relation between absorption at vertical incidence and absorption over an 800-km oblique-incidence path was studied by measurements on frequencies of 5.455 and 9.16 mc respectively. The observed results differ from those predicted by Martyn's absorption theorem by an amount which is subject to a diurnal variation. Empirical formulas are derived for diurnal and seasonal variation of absorption.

621.396.11.029.64:535.4 537

The Reflection of Electromagnetic Waves From a Rough Surface—H. Davies. (*Proc. IEE*, part IV, vol. 101, pp. 209–214; August, 1954. Digest, *ibid.*, part III, vol. 101, p. 118; March, 1954.) A statistical method is used to investigate scattering and reflection from perfectly conducting surfaces having random irregularities which are assumed to be not so large as to shield any part of the surface from the incident radiation. The method is used to study scattering of light by a disturbed water surface and to calculate the sea clutter in radar at centimeter wavelengths; the calculated results in the latter case are compared with observations reported previously [3257 of 1947 (Davies and Macfarlane)] and the limitations of the approximation are discussed.

621.396.812.029.62 538

Long-Distance U.S.W. Reception—E. A. Lauter and L. Klinker. (*NachrTech.*, vol. 4, pp. 242–247, 271; June, 1954.) Records of field-strength measurements made at Kühlungsborn since 1951 are analyzed in respect of diurnal variations of mean level, sudden changes in level, and rapid fluctuations. Data cover transmissions in the 87.5-100-mc band over two 200-km land paths and one 180-km sea path, and hourly records of some 30 transmissions over distances 50–500 km. Refraction theory accounts satisfactorily for propagation over distances up to three times the optical range; partial reflections can give field strengths below the horizon only 20–30 db below the level for free-space propagation; relatively high mean field strength at great distances during unstable atmospheric conditions is ascribed to scattering processes.

621.396.812.3 539

Spaced-Receiver Experiments on Radio Waves of 19-km Wavelength—J. Harwood. (*Proc. IEE*, part IV, vol. 101, pp. 183–186; August, 1954. Digest, *ibid.*, part III, vol. 101, pp. 113–115; March, 1954.) Signals from GBR were received at a distance of about 100 km at two spaced sites. Fading records of the wave reflected from the ionosphere indicated the presence of a randomly varying component superposed on the steady amplitude. The correlation between the records at the two receiving points was about 0.8 for a spacing of 15 km and 0.35 for a spacing of 40 km. The time variations of the random component were consistent with reflection from scattering centers in the ionosphere moving with random velocities having an rms value of about 8 m/s in the line of sight, or drifting steadily with a velocity of about 100 m/s in a direction predominantly east-west.

RECEPTION

621.376.332:621.396.822 540

The Statistical Properties of the Output of a Frequency Sensitive Device—G. R. Arthur. (*Jour. Appl. Phys.*, vol. 25, pp. 1185–1195; September, 1954.) Analysis is presented for a frequency discriminator with an input consisting of a narrow-band spectrum with random characteristics. For the case when the output is not filtered, a solution is found in which the output is given as the difference between two independent random quantities; the characteristic function method is used. For the case when the discriminator is followed by a filter, a solution is found by determining the predominant moments of the output probability density from a series expansion. This method shows that the Gaussian density is approached

when the output filter has a very narrow pass band.

621.396.62 541
Ratio Squarer—L. R. Kahn. (Proc. I.R.E., vol. 42, p. 1704; November, 1954.) Analysis indicates that in order to optimize the combined output signal/noise ratio in a diversity system in which the signals are added linearly and the noise according to root mean squares, the ratio of the signal levels should be squared before combination. A block diagram of a circuit used in the U. S. Army AN/FGC-29 equipment for obtaining the ratio squarer selection curve in a two-receiver system is shown.

621.396.621:621.314.7 542
Transistor Circuit—W. G. Walter and K. Walter. (*Wireless World*, vol. 60, p. 595; December, 1954.) The circuit is given of a highly selective receiver using a single junction transistor and a Ge diode. The signal drive is on the emitter and the rectified signal is derived from a coil in the base circuit.

STATIONS AND COMMUNICATION SYSTEMS

621.376:621.396.822 543
Theoretical Limits for Transmission in the case of High Noise Level for Different Continuous and Coded Modulation Systems—F. de Jager. (*Onde élect.*, vol. 34, pp. 675–682; August/September, 1954.) Using optimum bandwidth, the minimum power necessary for transmitting a sinusoidal signal to give a signal/noise ratio of 30–40 db at the receiver output is roughly the same for FM, PPM, binary-code PM and delta modulation systems, the noise level in the transmission channel being such that the signal/noise ratio obtainable with ssb modulation using the same power is 17–20 db. The required minimum power is a few db higher if pulses are modulated at hf. When the transmission chain comprises many sections the necessary power per section is much higher for continuous modulation systems such as FM and AM, but only a few db higher for coded systems.

621.376.5:621.39 544
Mean [energy] Spectrum of a Train of Pulses of Generally Periodic and Identical Character but Displaced and Deformed in a Random Manner—R. Fortet. (*Onde elect.*, vol. 34, pp. 683–687; August/September, 1954.) General formulas are derived, particular cases having been dealt with by Macfarlane (217 of 1950). The spectrum comprises: (a) lines at harmonics of the theoretical pulse repetition frequency, the amplitudes being dependent on the deformation and displacement of the pulses; (b) a continuous spectrum dependent on both deformation and displacement; this disappears if the pulse train is undisturbed. Thus the ratio of the total energies of (a) and (b) is effectively a signal/noise ratio.

621.396.41 545
A New Frequency-Division Multiplex Radiotelegraphy System (MADFAS)—A. Niutta. (*Poste e Telecomun.*, vol. 22, pp. 171–178; April, 1954.) Description of a frequency-shift system designed for high frequency stability and low frequency-band occupancy. The signal is generated in the af band and raised to radio frequency by successive modulations. A pilot frequency is transmitted for automatic control of the receiver oscillator. Tests made between Rome and Amsterdam gave satisfactory results.

621.396.712.3 546
New Unit-Construction Studio Equipment—G. Schadwinkel. (*Tech. Hausmitt. Nordw-Dtsch. Rdfunks*, vol. 6, nos. 5/6, pp. 137–140; 1954.) Details are given of the microphone supply unit Type N52a and the adjustable bandpass filter Type W75k.

621.396.933 547
A Single-Sideband Controlled-Carrier System for Aircraft Communication—G. W. Barnes. (*Proc. IEE*, part III, vol. 101, pp. 399–400; November, 1954.) Discussion on 2512 of 1954.

621.376.:621.39 548
Modulation Theory [Book Review]—H. S. Black. Publishers: Van Nostrand, New York and London, 1953, 363 pp., $8.75. (*Science*, vol. 120, p. 342; August 27, 1954.) The subject is considered in the light of information theory, with primary emphasis on the systems aspect of modulation. The material presented has been used in the communications development training program of the Bell Telephone Laboratories.

SUBSIDIARY APPARATUS

621–526 549
Analysis of Systems involving Difference-Differential Equations—T. M. Burford. (*Jour. Appl. Phys.*, vol. 25, pp. 1145–1148; September, 1954.) A method alternative to that presented by Chu (1185 of 1954) for analysis of control systems.

621–526:621.372.55 550
Compensation Networks for Carrier-Frequency Servomechanisms—P. Bonnet. (*Onde élect.*, vol. 34, pp. 688–699 and 812–818; August–October, 1954.) Methods are described for determining the transfer function and for the synthesis of gain- and phase-correction networks, particularly symmetrical bridged-T and parallel-T networks.

621.311.6:621.396.712 551
Power Supply for Automatically Controlled Transmitters—H. Kreutzträger. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 6, pp. 133–136; 1954.) Descriptions are given of the normal arrangements for taking power from the public supply and of emergency arrangements using batteries or Diesel generators.

621.314.634.3.012 552
Selenium Rectifier Instantaneous Characteristics—G. F. Pittman, Jr. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 73, pp. 45–49; March, 1954. Digest, *Elec. Eng.* (*N.Y.*), vol. 73, p. 719; August, 1954.) Current/voltage characteristics obtained by applying a rectified sinusoidal voltage to a Se cell in the reverse direction have been recorded. A current in the forward direction flows during a large part of the cycle. A convenient equivalent circuit to explain this result comprises nonlinear capacitance and nonlinear resistance in parallel. Both resistance and capacitance decrease with increasing reverse voltage and are relatively independent of the frequency of the supply voltage over a wide range.

621.396.63 553
Signal-Operated Switching—R. Selby. (*Wireless World*, vol. 60, pp. 613–615; December, 1954.) Description of a fault-warning system operated by cessation of an audio signal and used as a program-failure alarm. The circuit is based on a thyratron tube.

TELEVISION AND PHOTOTELEGRAPHY

621.397.26:621.396.65 554
The U.H.F. Radio-Link Equipment Type-RVG 904 for relaying Television Signals—C. Boden. (*NachrTech.*, vol. 4, pp. 267–271; June, 1954.) Review of the transmitter and receiver circuits and the antenna system of the equipment [3693 of 1954 (Mansfeld)], illustrating problems of design. A coaxial coupling system is used; the frequency-conversion stage at the input has a parallel-fed oscillator and push-pull IF output; the transmitter amplifier comprises four stages in cascade. Matching arrangements are outlined and operating characteristics are shown. Details are given of an uhf discriminator connected to the antenna feeder for monitoring purposes.

621.397.26:621.397.62:535.88 555
Large-Screen Projection Television: Demonstration at the Palais de Chaillot of News Item televised from Tours—R. Aubert and M. Sollima. (*Rev. gén. élect.*, vol. 63, pp. 533–535; September, 1954.) The demonstration was presented in May, 1954, to mark the occasion of the sixtieth anniversary of the École supérieure d'Électricité and of the Tours Week; it constituted the first radio relay of a televised news item by the French broadcasting organization. The existing Paris-Tours link was used in reverse. The relay stages and equipment used are indicated briefly; the pickup tube was a supericonoscope of French manufacture, and the projection system was of Schmidt type.

621.397.5 556
Reflections on the Future Development of Television—F. Schröter. (*Ricerca sci.*, vol. 24, pp. 1613–1640; August, 1954. In French.) A survey of problems involved, including program distribution, simplification of synchronizing-pulse scheme, collective reception, transmission of picture and sound on common carrier, and reduction of receiver costs. The imminence of color television in Europe enhances the need for unifying standards. Advantages of image storage at the receiver are discussed and various methods of compressing signal bandwidth are compared.

621.397.5:535.623 557
The Electronic Colour-Television System developed by the Société nouvelle de l'Outillage R.B.V. and the Société La Radio-Industrie—L. Kohn. (*Rev. gén. élect.*, vol. 63, pp. 527–532; September, 1954.) Report of developments in France, where efforts have been directed to devising a system simpler than that of the N.T.S.C. Using the whole 10.4 mc of the 819-line video channel, the first 8 mc is shared by the red and green signals in sequence, while the blue signal is transmitted continuously in a single sideband of width 1 mc on the outer side of a subcarrier at 9.4 mc. For convenience, 810 lines are used instead of 819, and the red and green signals alternate in line sequence, so that the red lines of one frame are superposed exactly on the green lines of the preceding one, i.e. there is no interlacing. Pickup arrangements for color film and for direct viewing are outlined, using three photocells and three image orthicons respectively, with optical splitting devices. Possible reception arrangements include (a) a three-tube projection system retaining the alternating red and green and continuous blue signals, (b) a single tube of chromatron type, involving sequential red, green and blue operation, and (c) use of an R.C.A. type three-gun color-dot tube, which again permits operation with a continuous blue signal.

621.397.5:535.623 558
Compatible Colour-Television Systems—É. Labin. (*Rev. gén. élect.*, vol. 63, pp. 516–526; September, 1954.) Conditions for compatibility between monochrome and color signals are considered and two particular systems are discussed in detail, that of the N.T.S.C. developed in the U.S.A., and that of Valensi, investigated in France. Valensi's system uses at the transmitter a special coding tube, having a screen with areas of different opacity corresponding to the 30 divisions of the color triangle recognized by the C.I.E., to produce a single-component chrominance signal from the three signals delivered by the trichromatic camera. A bandwidth of 1.2–1.5 mc is sufficient for this signal, which is placed on a subcarrier located near the edge of the video channel at a frequency which is an odd multiple of half the line frequency. At the receiver the three color signals are recovered by a decoding system in which the received chrominance signal is applied to three formatron tubes generating appropriate output functions determined from consideration of the nature of the C.I.E. coordinates. A construc-

tion in which the three formatron tubes are combined into one is described. Reception would require a variant of the synchronized-detection system combined with use of comb-characteristic filters. It would alternatively be possible to transmit Valensi-type signals by using the first 7 mc of the video channel for the luminance signal and the remaining 3 mc for the chrominance signal modulated on a sub-carrier, in which case synchronous detection would not be required. See also 557 above.

621.397.5:535.623 559
Colour Television—History, Present State and Future Development—E. Schwartz. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 6, pp. 105–126; 1954.) The developments culminating in the N.T.S.C. frequency-interlace system are described and the possibility of adapting this system for use in Europe is discussed.

621.397.6:535.623:621.385.832 560
Electronic Tubes for Colour Television—R. Juillerat. (*Rev. gén. élect.*, vol. 63, pp. 507–515; September, 1954.) A review of the different types of tube that have been developed both for the pickup and for the display of color-television pictures. Difficulties encountered in the design of suitable tubes, especially for reception, are having an important effect on the whole development of color television.

621.397.61:535.623 561
Problems of Television Cameras and Camera Tubes—L. H. Bedford. (*Jour. Brit. IRE*, vol. 14, pp. 464–474; October, 1954.) "Questions of sensitivity and signal-to-noise ratio are discussed. While the results are of general application, the subject is developed with particular reference to the image orthicon type of camera tube. The latter, in its standard form (the 3 inch image), has a considerable sensitivity advantage compared with other types of tube. For some purposes, particularly studio use, it is desirable to exchange some of this sensitivity for other properties, such as signal/noise ratio. In this case the balance of properties can be improved by increase in target area which leads to the large ($4\frac{1}{2}$-inch) image orthicon. A description is given of a new camera designed around the large image orthicon with maximum exploitation of the properties of the latter. Three approaches to the color television camera problem are described, and the repercussion of color on sensitivity and signal/noise ratio is discussed."

621.397.61.001.4 562
Color Text Technique for TV Broadcasters—Wentworth. (*See* 511.)

621.397.62 563
Flywheel Synchronizing—W. T. Cocking. (*Wireless World*, vol. 60, pp. 519–522, 557–560 and 621–624; October–December, 1954.) The effect of noise and interference on the line synchronization of television receivers is explained, and the use of balanced and unbalanced afc circuits for synchronization is described.

621.397.62 564
Television Safety Precautions—E. G. Goodhew. (*Wireless World*, vol. 60, pp. 591–592; December, 1954.) Brief discussion of considerations involved in drafting British and international standards covering precautions against the usual electrical hazards as well as tube implosion and incidental X radiation.

621.397.62 565
Television Intermediate Frequencies—(*Wireless World*, vol. 60, pp. 582–583; December, 1954.) A note on the considerations leading to the British Radio Equipment Manufacturers' adoption of 34.65 mc as vision IF for television receivers. A chart is presented showing the main possible forms of interference, for local-oscillator frequency above signal frequency.

621.397.62:621.386.842 566
Solid-State Image Intensifier—R. K. Orthuber and L. R. Ullery. (*Elec. Commun.*, vol. 31, pp. 198–201; September, 1954.) Reprint. See 3061 of 1954.

621.397.62.001.4 567
Measurements on Television Receivers: Part 5—Measurements on the Deflection System—O. Macek. (*Arch. tech Messen*, No. 224, pp. 205–208; September, 1954.) Part 4: 3634 of 1954.

621.397.621:621.375.232 568
Feedback I.F. Amplifiers for Television—H. S. Jewitt. (*Wireless World*, vol. 60, pp. 609–611; December, 1954.) Application of design principles described previously (1009 of 1954) to a vision IF amplifier with bandwidth about 3 mc; stray capacitance is an important feature. Performance figures are given for an experimental amplifier for a vision IF of 16 mc, the sound IF being 19.5 mc, the feedback is used in this case unaccompanied by transformer coupling or stagger tuning, though it can alternatively be used to give improvement in conjunction with these other features.

621.397.621.2:621.373.43 569
Differential Width Control for Television Line Scanning Circuits—K. G. Beauchamp. (*Electronic Eng.*, vol. 26, pp. 476–481; November, 1954.) A typical receiver line-scan circuit is studied to determine ways of varying the amplitude of the sawtooth current waveform without unduly affecting the extra-high voltage. An investigation is made of the design and performance of systems using a combination of series and parallel inductance connected to the output transformer, arranged so that the total inductance across the transformer windings remains constant.

621.397.82:621.396.61 570
Sine Wave all the Way—Blackie. (*See* 574.)

621.397.82.029.62 571
Band-III Television Interference—R. Davidson. (*Wireless World*, vol. 60, p. 594; December, 1954.) A note suggesting that the problem is less serious than indicated by Strafford (3708 of 1954).

621.397.828.029.62 572
Suppression of Interference from Motor Vehicles—W. Scholz. (*Elektrotech. Z., Edn A*, vol. 75, pp. 504–506; August 1, 1954.) Results of measurements, presented graphically, show that 95 per cent of cars (excluding diesel-driven vehicles) and 99 per cent of motor-cycles produce an interference field-strength $>500\ \mu v/m$ at 10 m distance at a frequency of 200 mc. This was measured using a receiver with a ± 100-kc passband in conjunction with a tuned dipole mounted 3 m above ground. The field-strength indication corresponding to 500 $\mu v/m$ would be 50 $\mu v/m$ on British or U.S.A. equipment. The use of built-in suppressors is recommended, e.g. resistors in spark plugs, as these have the advantage of being close to the source of interference and also of being inseparable from the engine component. Other suppressors are mentioned.

TRANSMISSION

621.396.61:621.376.4 573
The Circuit Development of the Ampliphase Broadcasting Transmitter—T. H. Price. (*Proc. IEE*, part III, vol. 101, pp. 391–398; November, 1954.) Two circuit arrangements are described for obtaining an amplitude-modulated output by differentially controlling the phase of two rf generators of substantially constant voltage operating with a common load; these are (a) the quarter-wave or constant-current ampliphase system, and (b) the push-pull or constant-voltage ampliphase system. Typical performance figures are quoted; these are equal to those for the usual types of transmitter circuit. Such designs are claimed to be highly economical as regards equipment and tubes.

621.396.61:621.397.82 574
Sine-Wave all the Way—L. Blackie. (*Short Wave Mag.*, vol. 12, pp. 312–315; August, 1954.) A discussion of the operation and efficiency of the final rf amplifier in amateur transmitters indicates that little reduction of available power is caused by changing from Class-C to Class-B, Class-AB2, or even Class-AB1 operation, while the corresponding reduction of harmonic output, and hence of interference with television, is considerable.

621.396.61:621.398 575
Pulse Transmitter for Rocket Research—D. G. Mazur. (*Electronics*, vol. 27, pp. 164–167; November, 1954.) The 15-channel AN/DKT transmitter installed in Aerobee rockets is described. The normal sampling rate in each channel is 312.5 per second, the input range is 0–5 v, the rf pulse width is 3 μs, the frequency 227 mc, the peak power output 10 w and the channel deflection range 150 μs. Provision can be made for a higher sampling rate in one channel by disabling three others.

TUBES AND THERMIONICS

621.314.63+621.314.7]:537.311.33 576
Semiconductors—[*Elec. Rev.* (*London*), vol. 155, pp. 307–311; August 27, 1954.] An account of work in progress at the new G.E.C. Wembley laboratories on the development of crystal tubes.

621.314.63:621.318.57 577
Transient Response of a *p-n* Junction—B. Lax and S. F. Neustradter. (*Jour. Appl. Phys.*, vol. 25, pp. 1148–1154; September, 1954.) Calculations are made of the diode current and the voltage across the junction over a period after reversing the applied voltage; typical values are assumed for the diode parameters. Results are shown graphically for various values of included series resistance, and are compared with those obtained by Kingston (2539 of 1954).

621.314.632:537.312.9 578
Influence of Mechanical Pressure on Barrier Height in Galena Rectifiers—J. N. Das and V. G. Bhide. (*Current Sci.*, vol. 23, pp. 185–186; June, 1954.) Measurements show that the barrier height of the galena/Pt-whisker contact rises from 1.275 to 1.830 and then falls to 0.320 v while the spread resistance falls from 82.10 to 15.61 and then rises to 67.55 Ω for pressures of 30, 110, and 150 g weight, using a whisker of radius 0.008 cm at the point of contact. Similar results were obtained with Al and C whiskers.

621.314.7 579
Theory of Diffusion-Type and Drift-Type Transistors: Part 2—Frequency Characteristics—H. Krömer. (*Arch. elekt. Übertragung*, vol. 8, pp. 363–369; August, 1954.) The quadripole admittances of a junction transistor (part 1: 3389 of 1954) taking account of the boundary-layer capacitances are dependent on frequency; a limiting frequency is thus introduced which is dependent on the matching conditions. The inherent frequency limit set by alpha cut-off is higher for the drift type than the diffusion type, in the case of Ge by a factor of 8. A detailed analysis is given for the case of purely ohmic matching. Performances are compared under different operating conditions. In practice better use can be made of the inherent cut-off frequency in the drift type.

621.314.7:001.4 580
Nomenclature of Transistor Parameters—H. Mette. (*Funk u. Ton*, vol. 8, pp. 420–421; August, 1954.) A list is given of recommended German equivalents of English transistor terms.

621.314.7:537.311.33 581
Transistor Electronics: Imperfections,

Unipolar and Analog Transistors: Parts 2 and 3—W. Shockley. [*Proc. IRE (Australia)*, vol. 15, pp. 194–201 and 228–234; August and September, 1954.] Reprint, see 746 of 1953. Part 1: 3600 of 1954.

621.314.7:621.317.3 582
Stability Considerations in the Parameter Measurements of V.H.F. Point-Contact Transistors—Thomas. (*See* 491.)

621.314.7:621.372.5 583
Design Considerations of Junction Transistors at Higher Frequencies—H. Statz, E. A. Guillemin and R. A. Pucel. (PROC. I.R.E., vol. 42, pp. 1620–1628; November, 1954.) An accurate equivalent network for the junction transistor is developed based on a solution of the diffusion equation. The relation between gain and frequency is calculated and the upper limiting frequency for obtaining power gain is determined; from these results it is possible to determine the proper values of collector capacitance and base resistance in order to make full use of obtainable base width. Use of the formulas in the design of transistors to have a prescribed upper limiting frequency is illustrated by numerical examples.

621.383.2 584
Increased Light Sensitivity from Standard Gas Phototubes—E. O. Johnson. (*Rev. Sci. Instr.*, vol. 25, pp. 839–840; August, 1954.) A circuit is described by means of which current amplification as high as 10^8 can be obtained with a commercial photocell without producing instability. The mechanism involves secondary emission at the photocathode due to bombardment by ions produced by the photoelectrons. The current build-up is terminated periodically by firing a thyratron.

621.383.2 585
The Behaviour of Gas-Filled Photocells operating in the Region of Self-Maintained Discharge—W. Kluge and A. Schulz. (*Z. angew. Phys.*, vol. 6, pp. 364–370; August, 1954.) It was shown previously [3264 of 1953 (Kluge)] that self-maintained discharge currents in gas-filled photocells can be controlled by illumination of the cathode. An investigation is made of the upper limit which the discharge current can attain without causing structural modifications of the composite cathode. Reversible fatigue effects due to bombardment by positive ions are observed; these effects are very small with potassium hydride cathodes, hence reproducible values of photocurrent can be obtained.

621.385.004.15 586
Developments in Trustworthy-Valve Techniques—E. G. Rowe and P. Welch. (*Elec. Commun.*, vol. 31, pp. 172–188; September, 1954.) Special tests for reliability are discussed and a complete testing procedure is indicated. Designing for reliability is considered as a problem distinct from that of making ordinary commercial valves more robust. See also 1776 of 1952 (Rowe).

621.385.029.6 587
Measurement of the Most Important Parameters of Helix-Type Travelling-Wave Valves for Amplification and Efficiency Calculations—W. Klein. (*Arch. elekt. Übertragung*, vol. 8, pp. 404–410; September, 1954.) Methods are described for determining the required parameters from measurements on completed prototype tubes. The results are useful for assessing the accuracy of calculated characteristics. Cases of special interest are (a) power tubes with relatively large difference between beam velocity and phase velocity on the helix, and (b) tubes in which the helix characteristics are appreciably modified by the dielectric supports.

621.385.029.62/.63:621.396.822 588
Minimum Noise Figure of Traveling-Wave Tubes with Uniform Helices—J. R. Pierce and W. E. Danielson. (*Jour. Appl. Phys.*, vol. 25, pp. 1163–1165; September, 1954.) An expression for the minimum noise figure is derived using a theorem presented by Pierce (304 of February) in conjunction with analysis developed by Watkins (1470 of 1952). Assuming reasonable values for the tube parameters, the value of 6 db is obtained for the minimum noise figure.

621.385.029.63/.64 589
Helix-Coupled Traveling-Wave Tubes—P. D. Lacy. (*Electronics*, vol. 27, pp. 132–135; November, 1954.) Two amplifier tubes with coupling helixes outside the vacuum envelope are described. Both operate over the range 2–4 kmc. One has a gain of 35 db, output 20 mw and noise figure of the order of 20 db; the other has a gain of 30 db, output 1 w and noise figure <30 db. The coupling mechanism is that of spatial beating [30 of 1950 (Krasnushkin and Khokhlov)].

621.385.029.63/.64 590
A Wide-Band Voltage-Tunable Oscillator—J. W. Sullivan. (PROC. I.R.E., vol. 42, pp. 1658–1665; November, 1954.) An experimental backward-wave tube has a bifilar helix in which matching is facilitated by forming one winding as a thread on a rod and the other as a wire insulated from the rod; the structure is essentially a coaxial line wound into a helix. Oscillations were obtained with frequency ranging from 2.6 to 13.3 kmc as the beam voltage was increased from 40 v to 3 kv, using appropriate magnetic focusing. Power outputs of 5–50 mw were obtained.

621.385.029.63/.64 591
Traveling-Wave Amplifiers and Backward-Wave Oscillators—M. Muller. (PROC. I.R.E., vol. 42, pp. 1651–1658; November, 1954.) Analysis is presented in a unified form for the travelling-wave tube, the magnetron amplifier and the corresponding backward-wave oscillators. Formulas are derived for amplifier gain, oscillator starting current and tuning parameters.

621.385.029.63/.64 592
Rippled Wall and Rippled Stream Amplifiers—C. K. Birdsall. (PROC. I.R.E., vol. 42, pp. 1628–1636; November, 1954.) Velocity-modulation tubes are discussed having space-periodic variations of diameter of drift tube or beam or both. The periodic length is a quarter of the plasma wavelength. Gain is produced by reinforcing the simple harmonic motion of the electrons in the beam; formulas are derived giving its value for various cases, including the use of velocity jumps. Design curves are presented. An experimental rippled-wall tube for the frequency range 1–4 kmcs described. See also 2498 of 1953 (Birdsall and Whinnery).

621.385.029.63/.64 593
Power Flow and Equivalent Circuits of Travelling-Wave Tubes—C. C. Wang, J. R. Pierce and P. K. Tien. (PROC. I.R.E., vol. 42, pp. 1701–1703; November, 1954.)

621.385.029.63/.64 594
A Computational Contribution to the Linear Theory of the Travelling-Wave Valve with Low Gain—D. Weber. (*Arch. elekt. Übertragung*, vol. 8, pp. 341–345; August, 1954.) For predicting low-gain operation the initial amplitude and phase of all three rf components must be known. These are calculated neglecting attenuation of the delay line and space-charge forces. A table of numerical values of characteristic parameters is given including a correction to those published earlier [3754 of 1953 (Schnitger and Weber)]. By means of a graphical method outlined initial conditions can be determined with sufficient accuracy in many cases without an exact construction. The results also apply to the carcinotron.

621.385.032.216 595
Cathode Parasitic Impedance—M. Berthaud. (*Onde élect.*, vol. 34, pp. 663–673; August/September, 1954.) Detailed discussion giving results of measurements made on commercial tubes at different cathode temperatures and of investigations of the formation of interface layers. The indications are that the whole of the cathode contributes to the increased impedance. Summarized version abstracted in 1620 of 1954.

621.385.032.216 596
Studies of the Interface Layer in Oxide Cathodes—L. S. Nergaard and R. M. Matheson. (*RCA Rev.*, vol. 15, pp. 335–361; September, 1954.) The complex nature of the interface layer is discussed and an experimental investigation is reported of the interface behavior of $2BaO.SiO_2$ which is formed by reaction between the BaO of the coating and Si diffusing out of the Ni base. No definite conclusions are drawn. Measurements of interface resistance should be made at normal-operation cathode current, otherwise the effect of the tube's previous history may be obscured.

621.385.032.216 597
Thermionic Emission and Electron Diffraction from Thin Films of Barium Oxide—P. N. Russell and A. S. Eisenstein. (*Jour. Appl. Phys.*, vol. 25, pp. 954–961; August, 1954.) The structure of and thermionic emission from evaporated BaO films on a Ni base were examined simultaneously, using a specially developed sealed-off electron-diffraction tube with fluorescent screen. The thermionic activity increased with film thickness up to about 20 monolayers and then remained constant up to 50 monolayers. The effect of heat treatment was examined. The emission density is comparable to that for sprayed oxide coatings, which are much thicker; the significance of this in relation to the emission mechanism is discussed.

621.385.032.216:621.317.331 598
Sensitive Apparatus for the Measurement of Cathode Parasitic Resistance—P. Sevin. (*Onde élect.*, vol. 34, pp. 673–674; August/September, 1954.) Voltages at 6 mc and 10 kc are modulated in phase quadrature by a 200-cps signal and applied simultaneously to the tube grid. A resistance across a 6-mc rejector circuit in the cathode lead is adjusted until modulation in the anode circuit is suppressed. Using an amplifier with a voltage gain of 10^5, internal resistance below 5 Ω in the cathode can be detected.

621.385.032.216:621.396.822 599
Noise Mechanisms in Oxide-Coated Cathodes—A. van der Ziel. (*Physica*, vol. 20, pp. 327–336; June, 1954.) Various explanations of flicker noise in tubes are reviewed. Schottky's emission-center theory gives best agreement with observations in the case of W, Th-W and L-type cathodes. For oxide-coated cathodes, calculations based on Nergaard's concept of a depletion layer (1528 of 1953) give values of noise of the same order of magnitude as the observed values.

621.385.5:621.375.2.024 600
"Starved Amplifiers"—Kaufer. (*See* 385.)

621.385.832 601
Operation of C.R.T. Storage Devices—S. Winkler and S. Nozick. (*Electronics*, vol. 27, pp. 184–187; October, 1954.) Types suitable for different applications are listed, with notes useful for design purposes. Limitations are indicated.

Gulton abstracts 3

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New ultrasonic drill gives maximum driving energy with minimum power

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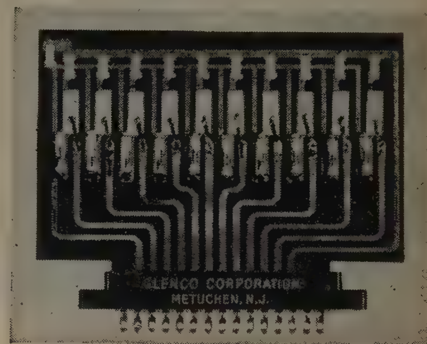

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Section Meetings

(Continued from page 122A)

SALT LAKE CITY

"Air Ionization and Medical Physics," by S. B. Hammond, University of Utah; December 9, 1954

SAN DIEGO

"Television in Color," by R. D. Kell, R.C.A. Labs.; October 19, 1954.

"Stereophonic Reproduction," by W. B. Snow, consulting engineer; November 2, 1954.

"Motorless Flight," by W. S. Ivans, Jr., Convair; December 7, 1954.

"Stabilization of Feedback Amplifiers," by Dr. D. C. Kalbfell, San Diego State College; January 4, 1955.

SCHENECTADY

"Particle Accelerators," by Dr. R. L. Kyhl General Electric Research Lab.; December 13, 1954

SYRACUSE

"Automatic Language Translation," by Dr L. E. Dostert, Georgetown University; December 2, 1954.

"Aims of the Professional Group," by Dr. W. R. G. Baker, "What is an Engineering Manager," by H. B. Fancher, "Low Frequency Filters in Hi-Fi," by N. S. Cromwell, all of General Electric Company, and "Modern Network Theory," by Dr. Norman Balabanian, Syracuse University; January 6, 1955.

TOLEDO

Round-table discussion on radio broadcasting in northwestern Ohio, by Chief Engineers of broadcast stations in that area; November 11, 1954.

"Educational Programming and Its Development," by M. W. Stahl, University of Toledo; December 9, 1954.

TORONTO

"Color Television," by P. A. Wigley; September 27, 1954.

"Recent Developments in High-Quality Sound Reproduction," by F. H. Slaymaker; October 25, 1954.

"Tube Application in TV Deflection Circuits," by S. F. Love; November 15, 1954.

"Wavestack Antenna," by G. M. MacKimmie; December 13, 1954.

TWIN CITIES

"Application of Electronic Computer Techniques to Heart Analysis," by Dr. O. H. Schmitt, University of Minnesota; December 7, 1954.

VANCOUVER

"Commercial TV Systems," by G. C. Chandler, Tru Vu Television Ltd.; November 15, 1954.

Field Trip to CBUT Television Studios; December 16, 1954.

WASHINGTON

"Time or Frequency Compression Expression of Speech," by Dr. W. L. Everitt, University of Illinois; January 10, 1955.

WINNIPEG

"Transistors in Audio," by C. Chaykowsky and "A Discussion on Accelerometers," by R. Wall, both students, University of Manitoba; December 1, 1954.

Guided tour of RCAF Telecom centre conducted by F/L D. Kyle, RCAF; December 15, 1954.

(Continued on page 136A)

reflectometer system for impedance measurement

Reflection coefficient or SWR measured instantly over wide frequency range

Direct, continuous swept-frequency oscilloscope presentation

Higher accuracy than slotted lines for single frequency measurement

Ends tedious point-by-point checking; system unaffected by amplitude variation

Ideal for fast production checks, system alignment, laboratory work checking waveguide components, antenna and rotary joint performance, etc.

Now *-hp-* offers a fast, accurate and *practical* reflectometer system that eliminates long hours of engineering previously required for microwave impedance measurements. The system provides direct meter readings of reflection coefficient, and is so simple to operate it can be used by non-technical personnel. An output is also available for oscilloscope or recorder presentation. The system eliminates the need for measuring forward and reverse power separately, and does away with tedious adjustments to correct for source amplitude variations. At present the equipment is available for X band operation only but components for other frequency ranges will be offered soon. The system includes several completely new *-hp-* instruments:

-hp- 416A Ratio Meter automatically combines forward and reverse signals and displays their ratio directly, irrespective of amplitude variations. Reflection coefficient may be read directly on the front panel meter. A signal at a rear terminal is provided to operate an oscilloscope or recorder. Model 416A contains an rf power monitor indicating proper power level. Obtainable accuracy for single frequency measurement is ±0.005 reflection coefficient; for swept frequency measurement, ±0.015 reflection coefficient.

Model 416A may also be used to measure SWR in connection with slotted lines. A reference voltage from the system power source applied to the ratio meter eliminates error due to amplitude variation.

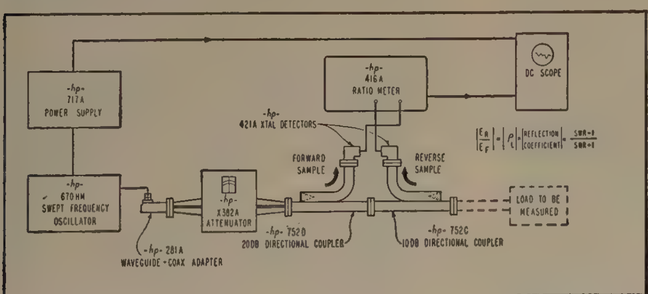

Figure 1. Reflectometer system. Swept rf power is provided by Oscillator. Directional Couplers sample forward and reverse power. Waveguide Detector Mounts terminating both Couplers demodulate power and present a 1,000 cps signal to Ratio Meter. Oscilloscope presents continuous visual study of reflection coefficient over the swept frequency range.

-hp- 670HM Swept Frequency Oscillator operates over a frequency range of 7 to 10 KMC. It may be manually tuned or motor driven to sweep any portion of this frequency band automatically. Sweep is at a velocity which is constant and sufficient to insure a clear trace on a long-persistence cathode ray oscilloscope. *-hp-* 670HM has a direct-reading frequency dial, and a waveguide-beyond-cutoff attenuator. It is normally grid modulated for sweeping, but also has reflector modulation for both fm and pulsed output. The oscillator requires an external power supply.

-hp- 717A Klystron Power Supply provides 800 to 1,000 volts beam voltage with regulation better than 0.1%, and 0 to 600 volts reflector voltage with regulation better than 0.05%. Model 717A also includes square wave modulation for both grid and repellor, as well as sine wave modulation for fm'ing. It is specifically designed as a companion instrument to *-hp-* 670HM.

-hp- 752 Series Directional Couplers are high directivity units consisting of two waveguides joined on their broad faces. Coupling is obtained through a series of precisely matched, graduated holes. In the system pictured, two couplers (with 10 db and 20 db directivity respectively) are joined back to back. Power flow is then as indicated in Fig. 1.

-hp- 421A Waveguide Detectors have square-law characteristics and demodulate the rf signal for use by *-hp-* 416A Ratio Meter.

Brief specifications of major instruments in the *-hp-* reflectometer system are given here. Complete specifications on all instruments in the system will be sent on request. For a comprehensive discussion of reflectometer measurements with these instruments, see *-hp-* Journal, Volume 6, Number 1-2, or write direct.

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Section Meetings

(Continued from page 130A)

SUBSECTIONS

AMARILLO-LUBBOCK

"Recent Developments in Transistor Materials and Devices," by Dr. G. K. Teal, Texas Instruments, Inc.; November 15, 1954.

"An Electronic Servo System for Guiding Telescopes," by Dr. J. H. Rush, Texas Technological College; December 9, 1954.

BERKSHIRE COUNTY

"UHF Television Trials and Tribulations," by Leon Podolsky, Greylock Broadcasting Company; November 12 and 13, 1954.

BUENAVENTURA

"Research Resources and Resourcefulness," by Dr. A. M. Zarem, Stanford Research Institute; December 9, 1954.

CENTRE COUNTY

"The A-C Network Calculator," by A. H. Forbes and P. E. Shields, Pennsylvania State Univ.; December 7, 1954.

MONMOUTH COUNTY

"Aircraft Flight Control and Navigation Systems Used by Man and Machine," by Edgar H. Fritze, Collins Radio Company; January 19, 1955.

PALO ALTO

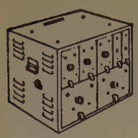
"Some Aspects of Human Communication," by Dr. Alexis Bavelas, Mass. Institute of Technology; January 13, 1955.

What to See at the Radio Engineering Show

(Continued from page 1A)

(Continued on page 140A)

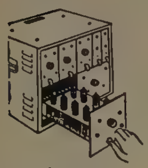
Electronic circuitry design and construction simplified by mechanical components to provide:



1

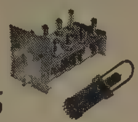
Circuitry sub-divided function by function into plug-in units.

ALDEN TERMINAL CARD SYSTEM



2

Plug in replacement spares in 30 seconds.



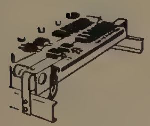
ALDEN BASIC CHASSIS & PLUG-IN PACKAGES

3

Tiny tell-tales spot trouble instantly.

ALDEN "TELL TALES"

4



All leads brought to single, accessible point of check, color coded and numbered so layman can make first-level check.

ALDEN BACK CONNECTORS

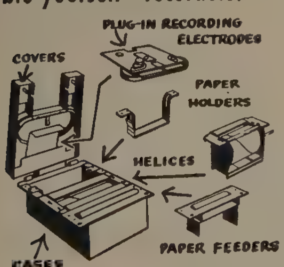
At the Alden Products Co. IRE Booth #185 & 187, or write Brockton, Mass.

Scientific recording now accomplished by Alfax Paper, where previously photography had seemed necessary.

New Alfax Paper — in which electricity is the ink — makes possible simple, uncomplicated recorders that operate unattended, get great amounts of related information compactly with high accuracy. Alfax paper is different-non-toxic, stable in storage, requires little current yet cannot overload, records at slow or high speeds, does not smudge or transfer.

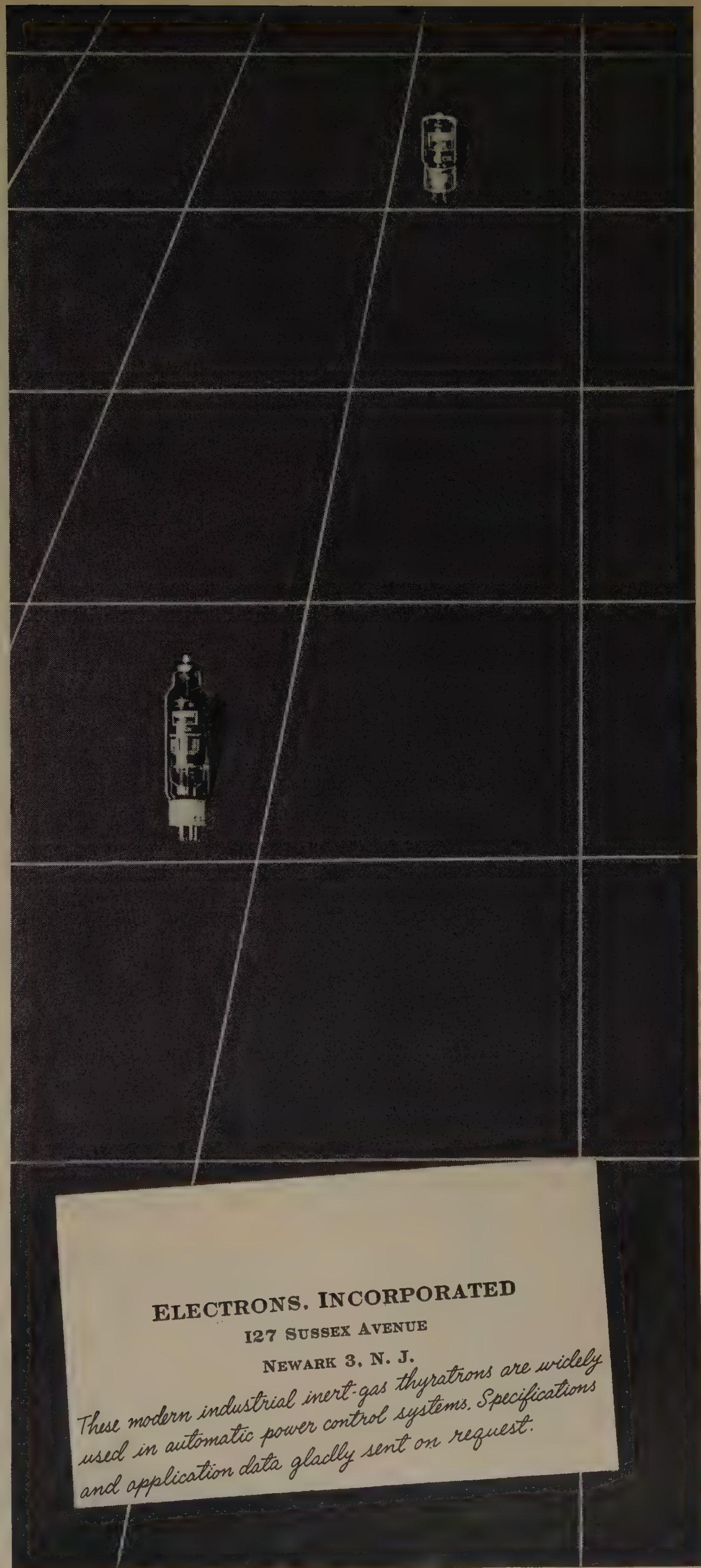
See the Alfax Paper and Engineering Co. IRE Show Booth No. 191 or write Westboro, Massachusetts.

Helix recording simplified by new "assemble yourself" recorders.



New horizons in graphic recording opened up by Alfax Paper — electricity is the ink — now can be explored quickly with these Alden adjusterless helix recorders and components.

See the Alden Electronic and Impulse Recording Equipment Co. IRE Booth No. 189 or write Westboro, Mass.



What to See at the Radio Engineering Show

(Continued from page 356A)

Wind Turbine Co., 591, 593 Components
*Broadband Curtain Antenna. *Corner Reflector Antenna. Towers and Masts.

Workshop Associates, 193-195, Television Ave.
See Gabriel Company

Yardney Electric Corp.
765 Airborne Ave.

Yardney Silvercel®, world's smallest, lightest, mightiest storage batteries—silver-zinc alkaline cells featuring constant output voltage and high current drain capacities—*1/10th AH and 250 AH high-rate, long-life Cells, *AN 3151 and *AN 3154 size-equivalent, high-capacity aircraft batteries.

Ziff-Davis Publishing Co., 839 Audio Ave.
See "Radio-Electronic-Engineering."

*Indicates new product

Show Hours: Monday 10 AM–10 PM
Tuesday 10 AM–10 PM
Wednesday 10 AM– 5 PM
Thursday 10 AM–10 PM

IRE People

D. B. Sinclair (J'30–A'33–M'38–SM'43–F'43) has been appointed Vice-President for Engineering of the General Radio Company. Dr. Sinclair joined the company in 1936, after receiving the Doctor of Science degree from M.I.T. He has served the company as Engineer, Assistant Chief Engineer, and, since 1949, as Chief Engineer.

D. B. Sinclair

During World War II, Dr. Sinclair was associated with the National Defense Research Committee program under Dr. Vannevar Bush, working on radar countermeasures and guided missiles. For his work overseas on the former project, he received the President's Certificate of Merit.

He is a Fellow of the American Institute of Electrical Engineers, a member of the American Association for the Advancement of Science, and of Sigma Xi, honorary scientific fraternity.

(Continued on page 360A)

IRE People

(Continued from page 358A)

A. M. Skellett (M'44–SM'50) has been named director of color television tube planning and development for Tung-Sol Electric Incorporated.

A. M. SKELLETT

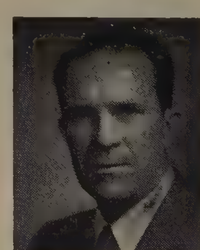
For the past 25 years he has been active in the electronics industry in research and administration, including 15 years on the technical staff of Bell Telephone Laboratories. He has also served the government as a consultant to the Research and Development Board of the Department of Defense. Among Dr. Skellett's patents is one for a radial beam tube which serves as an electronic commutator in military equipment and industrial instruments. Another, developed two years ago, is a magnetic tape pickup tube designed for recording equipment for high fidelity sound reproduction.

Dr. Skellett is a graduate of Washington University and received the A.B. degree in 1924 and the M.S., in 1927. Princeton University awarded him a Ph.D. degree in 1933. After graduation, he served two years on the University of Florida faculty, first as an instructor in electrical engineering and then as Assistant Professor of physics. Concurrent with his latter position, he was Chief Engineer of state-owned radio station WRUF in Gainesville, Fla.

He has written more than 30 articles and is a member of the American Physical Society and the American Astronomical Society.

❖

M. V. Kiebert, Jr. (A'31–M'38–SM'43) has recently been appointed Assistant to the Chief Engineer of Convair, a division of General Dynamics Corporation, Pomona, California.



M. V. KIEBERT

He was formerly Director of the Special Products Research Department of Bendix Aviation where he was in charge of the LOKI Missile Project and various other related Army Ordnance programs.

He has had long association with guided missile programs and related activities having been a member of the Aeronautical Board, Guided Missile Committee while serving as a Commander in charge of the Guided Missile Branch, Electronics Division, Bureau of Aeronautics, U. S. Navy.

(Continued on page 363A)

(*Continued from page 360A*)

During this period, Mr. Kiebert was responsible for technical design and coordination of the guidance and control problems of LARK, the initiation of the SPARROW project, major Navy computer projects, and Military Standardization programs in Telemetering. He has served on a number of panels and committees relating to missile, rocket and aeronautical electronics activities. He is a member of the Institute of the Aeronautical Sciences, the American Institute of Electrical Engineers, the American Ordnance Association, the American Rocket Society, the Society of Motion Picture and Television Engineers, the Franklin Institute, the Engineering Society of Detroit, and the Audio Engineering Society. He is also National Chairman of the Professional Group on Telemetering and Remote Control.

❖

R. C. Johnson (A'50), an engineer with the Alaskan Air Command, was recently presented the Meritorious Civilian Service award by Major General George R. Acheson, commander of the Alaskan Air Command. The award to Johnson was one of only two such awards ever presented in this overseas theater.

Mr. Johnson was presented the award for his chairmanship of the Traffic and Circuit Requirements Panel to the Alaskan Communications Study Group. This panel "established an integrated point-to-point communications circuit to tie in radar sites to military base areas in the Territory." Mr. Johnson recommended that all government and civil communication requirements be consolidated into a single, multichannel system to prevent duplication in engineering, manpower, equipment and circuits. The recommendation was then approved by Washington officials.

Prior to his arrival in Alaska in 1950, Mr. Johnson was employed by the Bell Telephone and Telegraph Company in Boston for 15 years. A native of Harmony, Maine, Johnson graduated from the University of Maine in 1930. He is a member of the American Institute of Electrical Engineers.

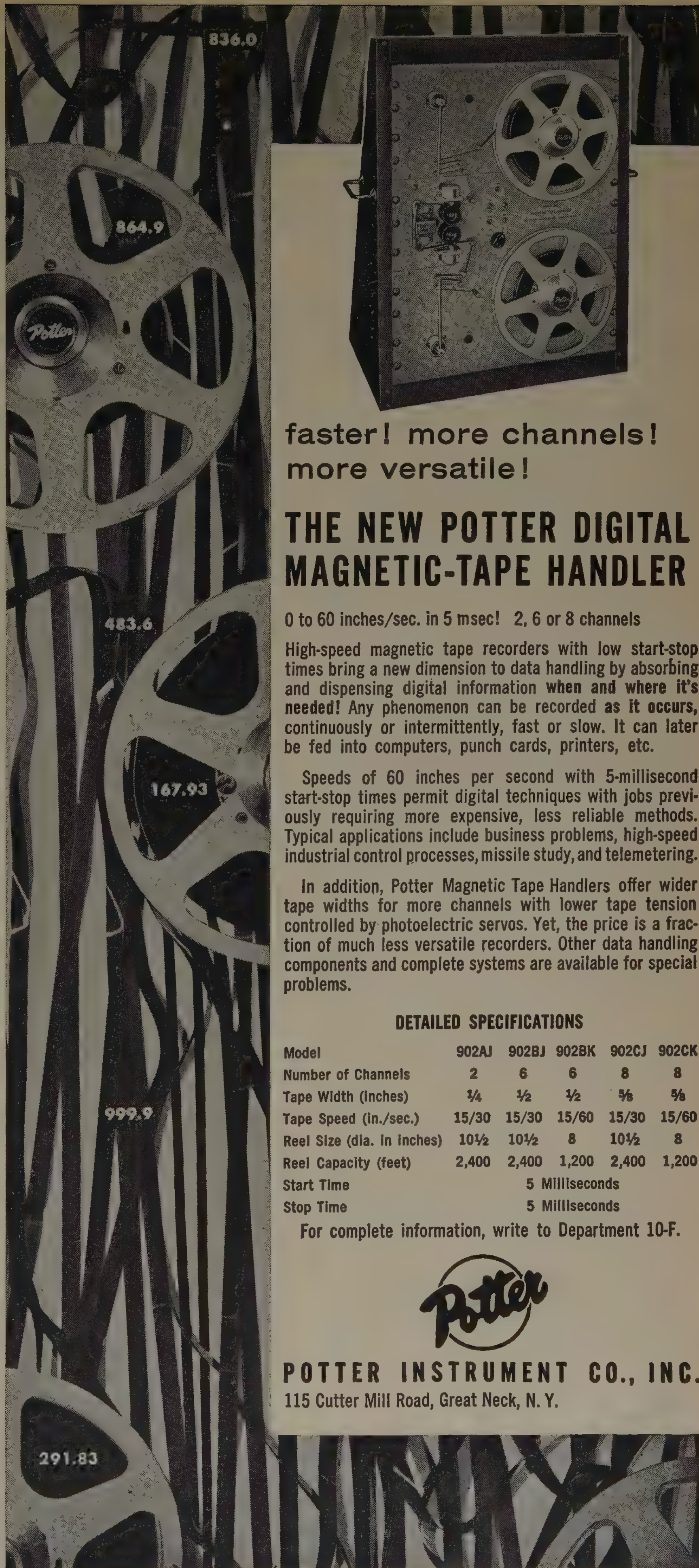
❖

J. E. Gall (A'37–VA'39) has been awarded the Navy's highest civilian award for his "outstanding" work in the development of improved electronic techniques and equipment. Mr. Gall is an electronics engineer with the Army Signal Corps in Washington.

The award was made by Assistant Secretary of the Navy James H. Smith and the citation reads:

"By your outstanding achievements in the development of improved electronic techniques and equipment while employed in the Naval Research Laboratory, Washington, D. C., you have made exceptionally significant contributions to the Navy. Your appreciation of human engineering

(*Continued on page 365A*)



LIGHT

RUGGED

traveling-wave tubes

Type SL24A—35 db minimum gain, 10 to 100 milliwatts output.

A series of tubes, as illustrated, is available, operating from 2 to 12 kilomegacycles—also complete amplifiers.

Note that the unit shown combines solenoid and tube with a total weight of only four pounds.

Write for detailed information.

An unusually comprehensive research facility is maintained at SLC for contract work in many phases of applied physics.

STANFORD LABORATORIES COMPANY

REDWOOD CITY CALIFORNIA

KLYSTRONS • PARTICLE ACCELERATORS

HENRY P. SEGEL'S

New Building Creates an Entirely New Concept in

MANUFACTURERS' REPRESENTATION

This architect's sketch illustrates the new headquarters building now under construction by the HENRY P. SEGEL COMPANY in Brookline (Boston Postal District) Massachusetts.

Here in these air-conditioned surroundings, industrial users and distributors will be able to view complete displays, inspect samples and obtain sales and engineering counsel regarding products made by the following manufacturers:

- BUD RADIO, INC.
- CAMPBELL INDUSTRIES, INC.
- CIRCUITRON, INC.
- CLAROSTAT MFG. CO., INC.
- CONNECTICUT TELEPHONE & ELECTRIC CORP.
- FREED TRANSFORMER CO., INC.
- GARDE MANUFACTURING COMPANY
- T V PRODUCTS COMPANY
- THE TURNER COMPANY

This modern building is strategically located in close proximity to all major routes in and out of Metropolitan Boston, the country's *leading* research and development center. A thorough survey of the New England market indicates that this structure will provide a long-needed facility to properly serve the requirements of the industry.

Manufacturers, wherever located, interested in having a type of representation in New England which will fulfill ALL requirements with vision and efficiency, plus warehousing facilities if desired, are invited to write for further information about this far-sighted and aggressive project.

HENRY P. SEGEL
Company Inc.

131 Newbury St., Boston 16, Mass.
Phone: KEnmore 6-3012, 6333, 9755

IRE People

(Continued from page 363A)

problems, complex electronic equipment and operational needs, has resulted in noteworthy pioneering accomplishment in the field of signal analysis. The new and revolutionary signal analysis equipment development on the basis of your endeavor has met with continuing success and has culminated in the production of valuable electronic equipment now in use by various components of the Armed Services. For these reasons, you are richly deserving of the Navy's Distinguished Civilian Service Award."

Mr. Gall, a native of Trenton, New Jersey, joined the Naval Research Laboratory in 1940, following work as an engineer with several radio manufacturing companies. He transferred to the Navy's Bureau of Aeronautics in 1950, and, in 1953, to the Army Signal Corps. Mr. Gall attended Purdue, Northwestern and George Washington universities.

❖

Vincent Salmon (SM'46), manager of Stanford Research Institute's sonics section, has been appointed editor of the "Journal of the Audio Engineering Society." He succeeds Lewis S. Goodfriend, now editor of "Noise Control" published by the Acoustical Society of America. The journal deals with advances in techniques and instruments for high fidelity recording and reproduction of sound, in studio acoustics, audio components and measuring gear, public address and stereophonic sound systems design and in electronic musical instruments.

Before joining SRI in 1949, Dr. Salmon headed research and development at Jensen Radio Manufacturing Company, where he worked on horn theory, multiple-source loudspeaker units and the design of devices used in radar countermeasures.

In 1946 Dr. Salmon received the biennial award of the Acoustical Society of America for the invention of a new family of horns. He is a Fellow of the AES and ASA.

❖

E. C. Ballentine (A'25–M'29–SM'43) has been promoted to Director of Engineering for KORCAD, RCA distributors for Korea. In his new position he is responsible for all engineering activities and policies of the new company. Mr. Ballentine is also serving as consulting engineer for several associated companies in Korea.

E. C. Ballentine

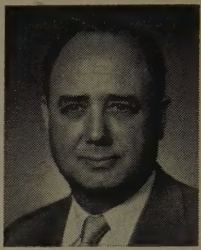
A graduate of the University of North Carolina with the B.S. degree in electrical engineering, he was, for a number of years, with General Electric Company.

(Continued on page 366A)

I R E People

(Continued from page 365A)

G. K. Teal, Assistant Vice-President, has been appointed to head the Research Division of Texas Instruments Incorporated. As Assistant Vice-President for Research, he now has charge of all research; the work involving many phases of electronics and geophysics. Previously he headed the Materials and Components Research Department, with primary responsibility for semiconductor research.



G. K. Teal

Dr. Teal was associated with Bell Telephone Laboratories for 22 years before coming to Texas Instruments in January of 1953. He has conducted research in photoelectricity, secondary emission, photoconductance, electron multipliers, silicon carbide varistors, heavy hydrogen, germanium and silicon rectifiers, microwave attenuators, borocarbon resistors, germanium and silicon single crystals, and transistors. He has some 46 patents granted or applied for as a result of his research in these fields.

A native Texan, Dr. Teal was born in Dallas and was graduated from Baylor University in 1927. He received the Master's and Doctor's degrees in chemistry from Brown University, and later was research associate in Prof. Harold C. Urey's laboratories at Columbia University. He is a member of the American Chemical Society, American Physical Society, the Texas Academy of Science, Sigma Xi, and is presently Secretary-Treasurer of the Dallas-Fort Worth section of the I.R.E.

❖

Servomechanisms, Incorporated has named **Robert Tate** (A'53) to the position of Sales and Service Manager for its Eastern Division. Prior to this appointment, Mr. Tate, who is well known in the field of aircraft instrumentation and control, was Contracts and Service Manager for Kearfott Company, Incorporated.

Robert Tate

Mr. Tate is a member of the Institute of Navigation and the American Ordnance Association.

❖

H. H. Newby (A'50–M'51), Chief Engineer for KAKE Radio since it went on the air seven years ago, has been confirmed as

(Continued on page 368A)

IRE People

(Continued from page 366A)

H. H. Newby

Chief Engineer for KAKE-TV, Wichita's first VHF station to begin telecasting operations. He has been in charge of engineering operations since last May.

Separate transmitting and studio plants are now under construction. A 50 kw RCA transmitter and 1,000 feet Parkersburg tower topped by a 12-bay RCA antenna are a part of the station's equipment. A three-quarter million dollar facility in the city houses operations. The main studio measures 60 by 80 feet and a second studio is 40 by 30 feet.

❖

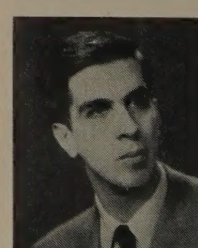
L. M. Orman (SM'52) has been made Chief of the Fire Control Division, Development and Proof Services, Aberdeen Proving Ground, Md. Colonel Orman's division is responsible for acceptance testing and development work on the Army's fire control systems. Last year he held the position of Head of Test Group OCAFF Bd #4, Fort Bliss, Texas. He is the author of "Electronic Navigation" and many articles in service journals on radar and related subjects.

L. M. Orman

Lt. Col. Orman holds the B.S. degree from the U. S. Military Academy and the M.S. degree from the Moore School of Electrical Engineering, University of Pennsylvania.

❖

Promotion of **R. L. Rod** (S'42–A'43–M'46–SM'50) to assistant to Vice-President of Bogue Electric Manufacturing Company has been announced. Mr. Rod will assist J. A. Herbst, Vice-President-Engineering, in the administration of engineering activities of the firm.



R. L. Rod

Previously, Mr. Rod was Assistant Director of Research and Development and headed the company's activities in the development of electronic devices. Before joining Bogue in 1951, he was associated with Melpar, Incorporated and Radiomarine Corporation of America. He has assisted in the development of radar, microwave communications systems and ultrasonic instruments.